Electronic technology

principal contents

FIVE SHILLINGS

FEBRUARY 1961 VOLUME 38 NUMBER 2

STORAGE TUBES

'ENGLISH ELECTRIC'

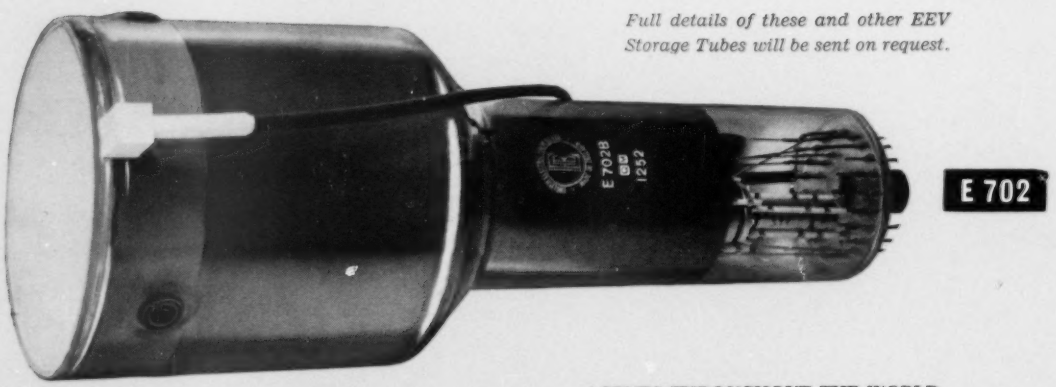
FIRST AGAIN IN BRITAIN

The English Electric Valve Co., consistent with their policy of keeping in the forefront with new electronic devices, have recently enlarged considerably their Storage Tube Division. Increased facilities and effort are being applied to the development and production of a number of Storage/Memory Tubes, two of which are illustrated, and enquiries are invited for special types to meet individual requirements.

E702 Direct-view Storage Tube has a useful screen diameter of 4 inches and information can be displayed at high brilliance (2350 foot-lamberts) for one or two minutes. The persistence can be controlled for 10 minutes or more, or the display can be completely erased in 300 milliseconds.

E704 Island Storage Tube will store about 500 bits of information and is designed for analogue and digital storage and for radar data handling applications, where it gives improved signal to noise ratio. The electrical input and output may be applied or taken out at the same or at different rates; the two processes may be simultaneous or sequential.

Full details of these and other EEV Storage Tubes will be sent on request.



AGENTS THROUGHOUT THE WORLD

ENGLISH ELECTRIC VALVE CO. LTD.

Chelmsford, England
Telephone: Chelmsford 3491

AP149

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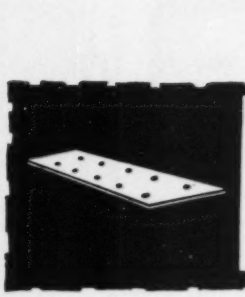
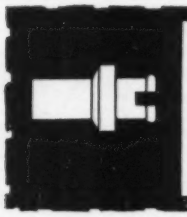
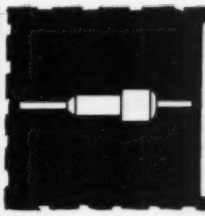


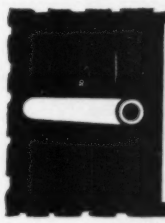
TERMINAL BOARDS



LUGS



MOULDED INSULATORS



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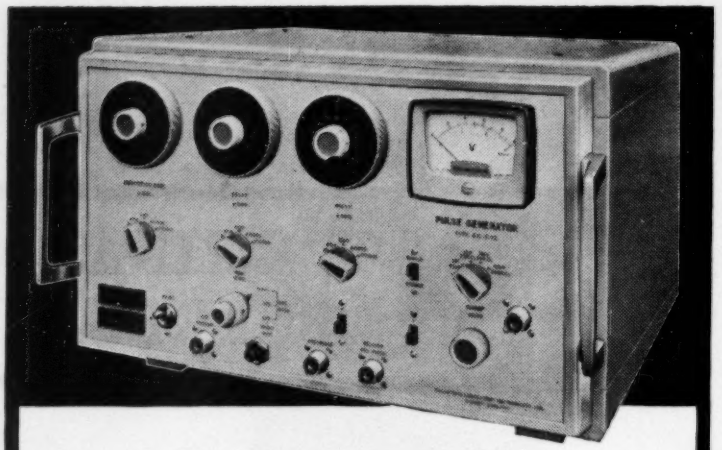
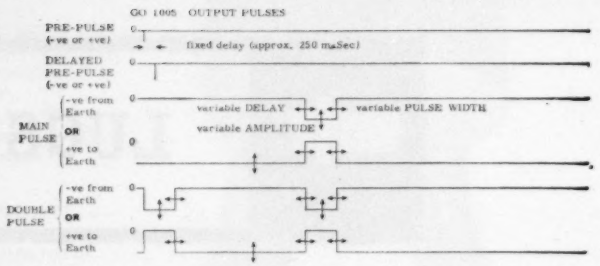
SOLARTRON

DECADE PULSE GENERATOR GO 1005

Decade Controls
Repetition frequency 10 c/s—1 Mc/s
Variable DELAY, PULSE DURATION and PULSE AMPLITUDE
Prepulse and Delayed Prepulse
Single Shot facility
External trigger up to 2 Mc/s

The extremely versatile Solartron Decade Pulse Generator GO 1005 provides single or double rectangular pulses of up to 100 volts amplitude, monitored by a built-in meter, at repetition rates from 10 c/s to 1 Mc/s.
The desired output conditions are easily and accurately set with illuminated concentric decade dials, permitting both coarse and fine adjustments of pulse-duration, repetition rate and pulse-delay. Extensive delay facilities are incorporated in the GO 1005 and triggering may be internal or external up to 2 Mc/s.





SPECIFICATION

Repetition frequency:	10 c/s—1 Mc/s (calibrated as repetition time 1 μsec—10^5 μsecs), continuously variable.
Frequency accuracy:	$\pm$ 5% from 10 c/s—100 kc/s, $\pm$ 10% outside these limits.
Pulse duration:	250 mμsec—100 msecs, accurate to $\pm$ 5% continuously variable. (Minimum pulse width 200 mμsecs).
Pulse amplitude:	0—100V, accurate to $\pm$ 3%; continuously variable. Polarity: —ve going from earth potential or +ve to earth potential.
Rise time: (-ve pulse)	20, 30, 60, 200 mμsecs for amplitude ranges 0—3, 10, 30, 100 volts respectively.
Double pulses:	Two identical main pulses, leading edge of the first coincides with prepulse.
Delay time:	(relative to Prepulse) 250 mμsec —100 msecs, to $\pm$ 5% accuracy, continuously variable.
Prepulse:	10 V amplitude (+ve or —ve), 150 mμsecs. duration.
Delayed Prepulse:	Fixed delay of $\pm$ 250 mμsecs after the Prepulse. (polarity same as prepulse).
Triggering:	Internal or external (up to 2 Mc/s) with trigger level control to select trigger point. Single shot facility.

Write now for full details to:
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Cox Lane, Chessington, Surrey.
Telephone: LOWer Hook 2150 Cables: SOLARTRON CHESSINGTON
Telex: 23842 SOLARTRON T.DIT. *A Member of the Firth Cleveland Group*

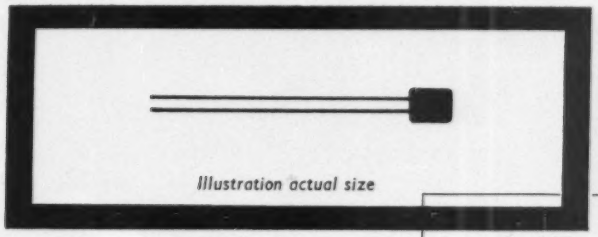

Illustration actual size

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silicon
VOLTAGE REGULATOR DEVICES

★ 150 mW Types KS30 to KS44	
Ref. Voltage range measured at 5mA.	3·3 to 12 volts
Tolerances on ref. voltage	5% & 10%
Max. Dissipation up to 75°C.	150 mW
Derating above 75°C.	2 mW/°C.
Operating temp. range	–40 to +150°C.

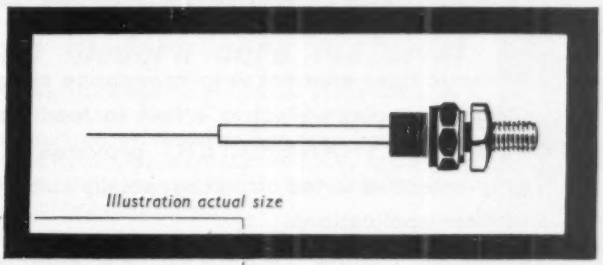

Illustration actual size

★ 8 Watt Types KR50 to KR60	
Ref. Voltage range measured at 50 mA.	15 to 100 volts
Tolerance on ref. voltage	5% & 10%
Max. Dissipation at 25°C. Ambient Temp. Mounted on $2\frac{1}{2}$" x $2\frac{1}{2}$" x $\frac{1}{16}$" Heat Sink	8 watts
Derating with Temp. above 25°C.	73 mW/°C.
Operating range	–40 to +135°C.

★ ***HERMETICALLY SEALED***

★ ***LOW ZENER IMPEDANCE***

★ ***RUGGED CONSTRUCTION***

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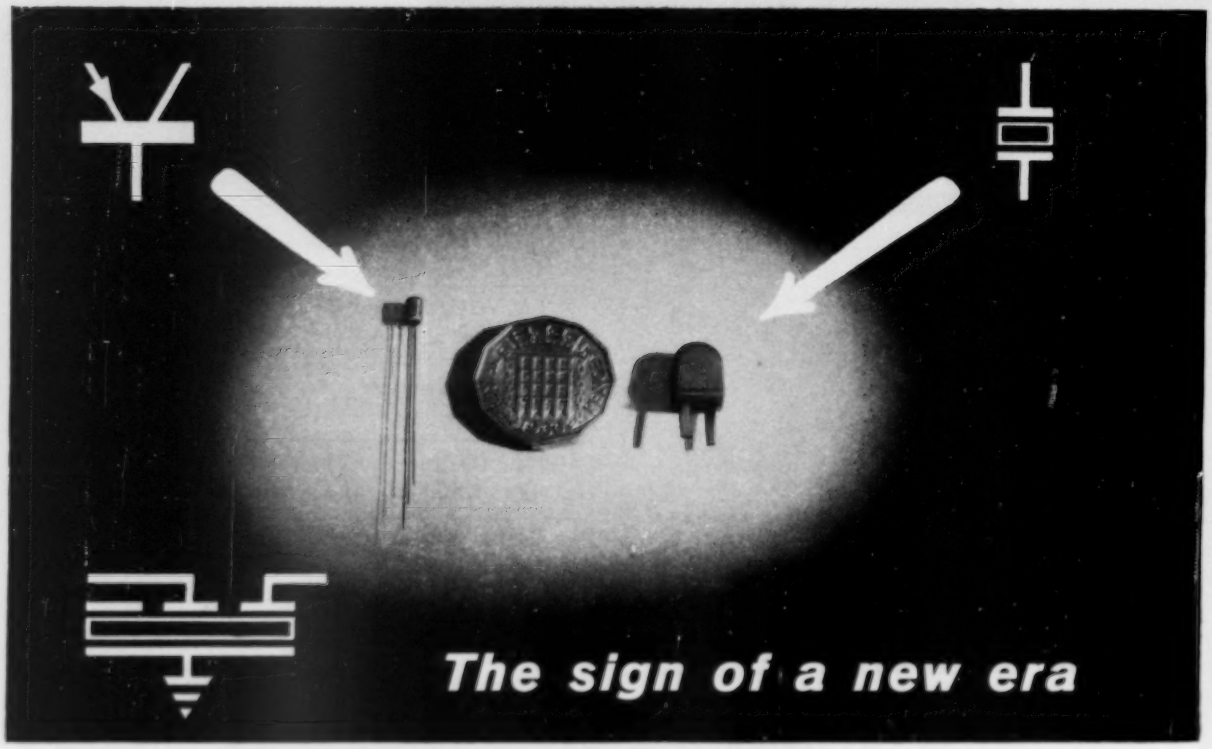


A ceramic filter element with impedance coupling properties! Utilizing the piezo-electric effect in lead zirconate titanate ceramic, the TRANSFILTER* provides a low-impedance highly-selective tuned circuit especially suitable for i.f. circuits and filter applications.

The Brush 'Clevite' TRANSFILTER combines ideally with transistors in the trend for miniaturization. Designed for use with printed circuits, the TRANSFILTER requires no initial or periodic frequency alignment and is unaffected by shocks exceeding 100 g or temperatures up to 200°C.

Write for the TRANSFILTER leaflet, which gives details of types available for 455, 465 and 500 kc/s operation in interstage coupler or emitter by-pass applications.

*Reg'd. Trade Mark

BRUSH
CRYSTAL CO. LTD.
CERAMICS

HYTHE · SOUTHAMPTON · HANTS · Phone: HYTHE 3031

ALPHASIL—*the modern core material*

The inset curves illustrate the superior magnetic properties of Alphasil cold-reduced grain oriented silicon steel over those of a typical hot-rolled grade (Ferrosil 80). Alphasil has a maximum permeability four times that of the hot-rolled transformer sheet and its core losses are approximately one-third. Initial and incremental permeability, stacking factor and ductility are considerably better than those of hot-rolled sheet.

Alphasil ·013" thick is produced in coil 30 inches wide, and can be supplied slit to narrower widths by arrangement.

TABLE OF WATTS LOSSES

	Frequency cycles/second	*Guaranteed max. total losses at B. Max. 15 Kilogauss*
ALPHASIL 44	50	·62 watts/lb.
ALPHASIL 40	50	·56 watts/lb.
ALPHASIL 37	50	·51 watts/lb.
ALPHASIL 33	50	·46 watts/lb.

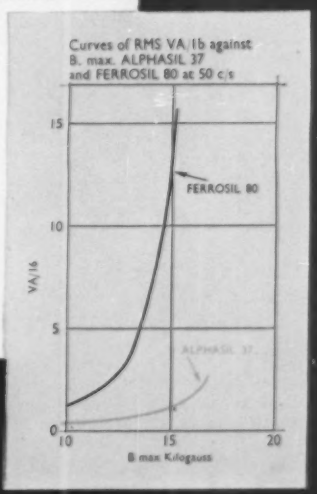
ABOVE—
A 4,000-lb. coil of 30" wide x ·013" thick, ready for despatch.

RIGHT—
Core-loss testing of Alphasil by the 'double-lap' Epstein method.

Thin Alphasil for high frequency work is also available in coil in ·004" thick, in widths up to 5½ inches, and in ·002" thick, in widths up to 4½ inches.

	Frequency cycles/second	*Guaranteed max. total losses*
ALPHASIL ·004HF	400	8·00 watts/lb. at B. Max 15 Kilogauss
ALPHASIL ·002HF	8,000	9·50 watts/lb. at B. Max 2 Kilogauss

Full technical data will be supplied on request



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2 tape speeds
(3 3/4″ and 7 1/2″ or
7 1/2″ and 15″ per second)
Can be used with or
without microphone preamplifier,
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Type F 100 M : multi-track (2 tracks : stereo)
Overall response curves F 100 and F 100 M :
15″ per second : 40 to 15,000 c/s ± 2 dB
7 1/2″ per second : 40 to 15,000 c/s + 2 −4 dB

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ampli de mélange et de contrôle.
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Courbes de réponse globales F 100 et F 100 M :
38 cm/s : 40 à 15 000 Hz ± 2 dB
19 cm/s : 40 à 15 000 Hz + 2 − 4 dB

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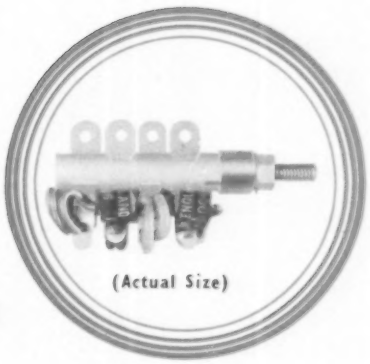

(Actual Size)

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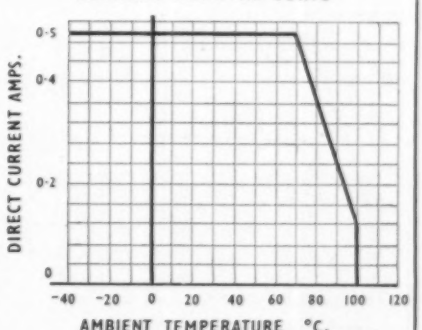
RATINGS (UP TO 2400 C/S)				
CODE	ABSOLUTE P.I.V. RATING	V r.m.s*	I.D.C. MAX.	TEMP.
IB000	50	35	0·5	70°C
IB001	100	70	0·5	70°C
IB003	200	140	0·5	70°C
IB006	400	280	0·5	70°C

* FOR DIRECT MAINS OPERATION DERATE BY A FACTOR OF 2.

CURRENT DERATING CURVE



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G. & E. Bradley Ltd

SEMI-CONDUCTOR DIVISION · Telephone Gladstone 0012
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for transistorised circuits

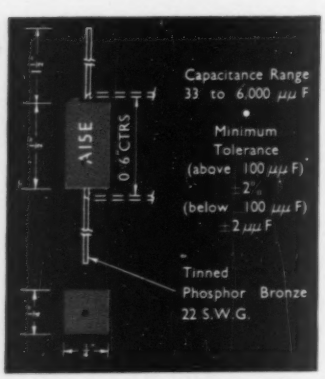
Something new in silvered mica capacitors — new in the Silver Star range!

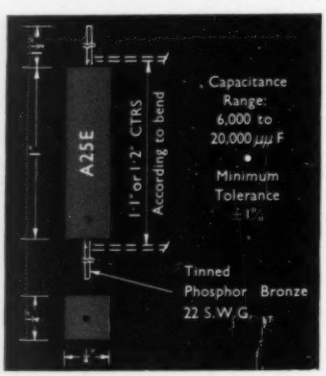
Designed specially for transistor circuitry, these miniature square-section capacitors are particularly useful when the component density is high.

And they are ideal for printed circuits, for they can be cemented to the circuit board — a big advantage when acceleration forces are high.

Two sizes (A15E and A25E) together cover the capacitance range from 33 to 20,000 μμF. For values above 6,000 μμF the capacitance tolerance can be as close as ±1 per cent.

- **Square section**
- **Miniature size**
- **Ideal for printed circuitry**
- **50 volt rating**
- **Fired construction for maximum long-term stability**
- **Synthetic resin encapsulated**
- **Ideal for "high g" applications**
- **Silver Star® quality**





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with this **NEW** G.E.C. valve

Here is our new low impedance double triode, the 6080 WA. In the first place, you save space with these two valves in one envelope. In the second place, here is a really fine, special quality double triode, the result of our 40 years of experience in manufacturing valves of all kinds. Full information on all Series Stabilisers, including the 6080 W A is available, as well as a detailed report on the design of Series Stabiliser circuits.

6080 W A

Maximum anode voltage	250V
Maximum cathode current	2 x 125mA
r_a	280+280Ω

The full range of series stabilisers cover these voltages: 80 mA at 250V to 1·5A at 2kV

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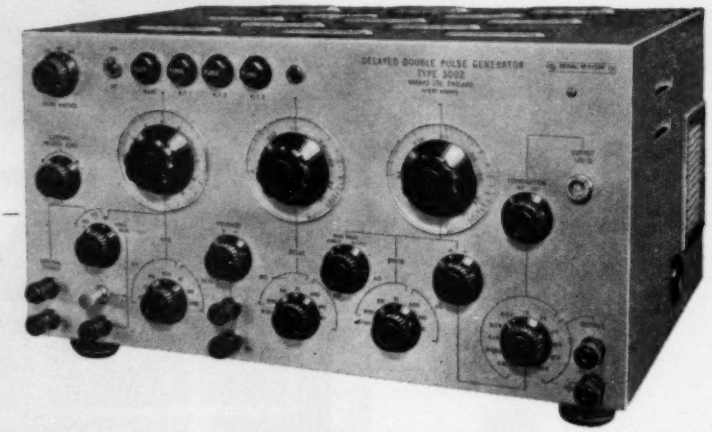
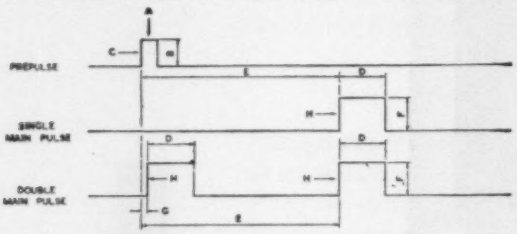
10,000,000 to 1

continuously variable RANGE of Square Pulse Width, Delay and Rate



- **SINGLE OR DOUBLE PULSES of either polarity.**
- **FAST RISE TIME—10 to 12 mμsec. independent of control settings over a range of 20mV. to 2V.**
- **AMPLITUDE—Accurately calibrated within 2%—20mV to 50V.**
- **NEGLIGIBLE RINGING, OVERSHOOT, JITTER and NO SAG.**
- **INTERNAL RATE GENERATOR 0·1c/s to 1Mc/s.**
- **EXTERNAL TRIGGERING up to 2·5Mc/s from any wave form.**
- **SINGLE PULSE or PAIR OF PULSES by push button.**

ALL THESE FEATURES make the NAGARD 5002 the most versatile PULSE GENERATOR available today



A. 0·2μSEC.
B. 20VOLTS +ve OR −ve
C. RISETIME 0·03μSEC
D. 0·2μSEC. TO 2SEC. CONTINUOUSLY VARIABLE ± 5% ± 0·02μSEC
E. 20mV TO 50V ± 2% IN 11 STEPS VARIABLE +ve OR −ve
G. APPROX. 0·04μSEC
H. RISETIME 10–12mμSEC. 20mV TO 2V. 5–50V DEPENDENT ON LOAD

REPETITION RATE CONTINUOUSLY VARIABLE 0·1c/s TO 1Mc/s ± 5% BY INTERNAL GENERATOR, OR TO 2·5Mc/s BY EXTERNAL TRIGGER SOURCE.

The diagram on the left represents its output

You can now visualise some of its many applications, for example:—

TRANSISTOR CIRCUITRY
The testing of response characteristics, particularly in computer applications.

COMPUTERS
And all devices depending on gating circuits can be tested for response times.

PULSE GROUP GENERATOR—TYPE 5101
This instrument added to the 5002 Pulse Generator produces groups of pulses which can be varied over a wide range of group rates and numbers of pulses per group—ask for Data Sheet No. 1060.

To get the best results use it with a NAGARD OSCILLOSCOPE

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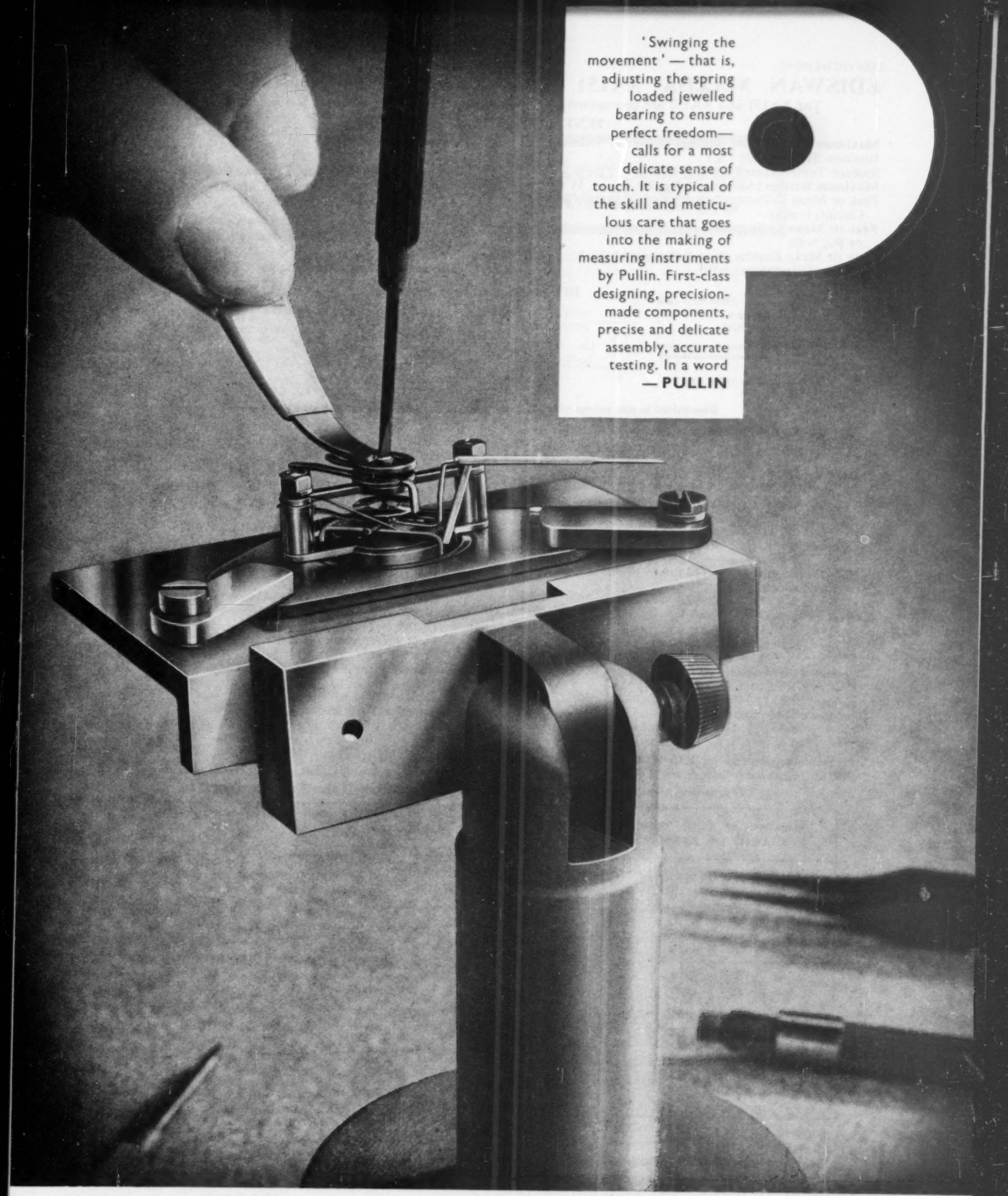

'Swinging the movement' — that is, adjusting the spring loaded jewelled bearing to ensure perfect freedom— calls for a most delicate sense of touch. It is typical of the skill and meticulous care that goes into the making of measuring instruments by Pullin. First-class designing, precision-made components, precise and delicate assembly, accurate testing. In a word **— PULLIN**

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EDISWAN MAZDA XA151 AND XA152 SWITCHING TRANSISTORS

The XA151 and XA152 are germanium pnp alloy junction transistors intended for switching applications.

TENTATIVE RATINGS AND DATA

Maximum Temperature Ratings (Absolute Values)

Junction Temperature (°C)	65
Storage Temperature (°C)	65

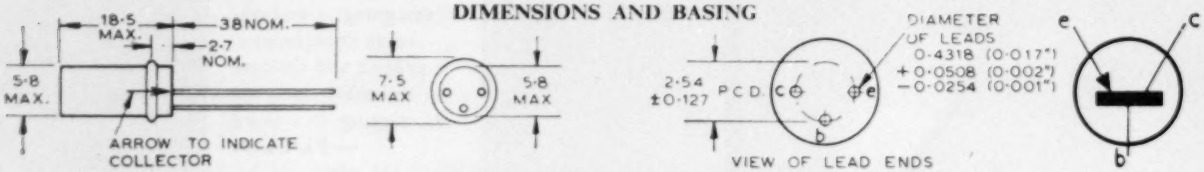
Maximum Ratings (Absolute Values for T_{amb} = 45°C)

Peak or Mean Collector/Base Voltage (Common Base Circuit) (volts)	−15
Peak or Mean Collector/Emitter Voltage ($V_{be} \geq$ 0·5V or R_{be} = 0)	−15
Peak or Mean Emitter/Base Voltage (volts)	−12
Collector Dissipation (mW)	66

General Characteristics

	T_{amb} (°C) 25	60
Maximum Collector/Base Leakage Current		
(V_{cb} = −15 V, Emitter open circuit) (μA)	—	−50
(V_{cb} = −15 V, V_{be} = +0·5 V) (μA)	−10	−50
Maximum Emitter/Base Leakage Current		
(V_{eb} = −12 V, Collector open circuit) (μA)	−10	—
Thermal Resistance in Free Air (°C/mW)	0·3	—

DIMENSIONS AND BASING



Dimensions in mm except where stated otherwise. Dimensions include insulated sleeve.

XA151 SWITCHING CHARACTERISTICS

	Typical Production Spread Min.	Av.	Max.
Static Current Amplification (d.c.)			
(V_{ce} = −0·25 V, I_c = −10 mA)	20	—	90
(V_{ce} = −0·25 V, I_c = −50 mA)	20	—	80
(V_{ce} = −0·25 V, I_c = −125 mA)	10	—	60
Collector/Emitter Saturation Voltage (volts)			
(I_c = −10 mA, I_b = −0·6 mA)	−0·04	—	−0·14
(I_c = −50 mA, I_b = −3 mA)	−0·06	—	−0·2
(I_c = −125 mA, I_b = −14 mA)	−0·06	—	−0·28
Base Resistance r_{bb}' (Ω)			
(V_{ce} = −6 V, I_e = −1 mA)	—	75	—
Cut-off Frequency f_1 * (Mc/s)			
(V_{ce} = −2 V, I_e = −10 mA)	3	—	—
Collector Depletion Capacity (pF)			
(V_{cb} = −6 V, I_e = 0)	—	11	16
Collector Time Factor τco (mμs)			
(V_{ce} = −0·25 to −4·25 V I_c = −10·85 to −0·85 mA)	—	30	70
Saturation Time Factor τs (μs)			
(I_c = −10·85 mA; I_b = −0·62 mA)	—	0·8	—
βs† (I_c = −10·85 mA; I_b = −0·62 mA)	—	27	—

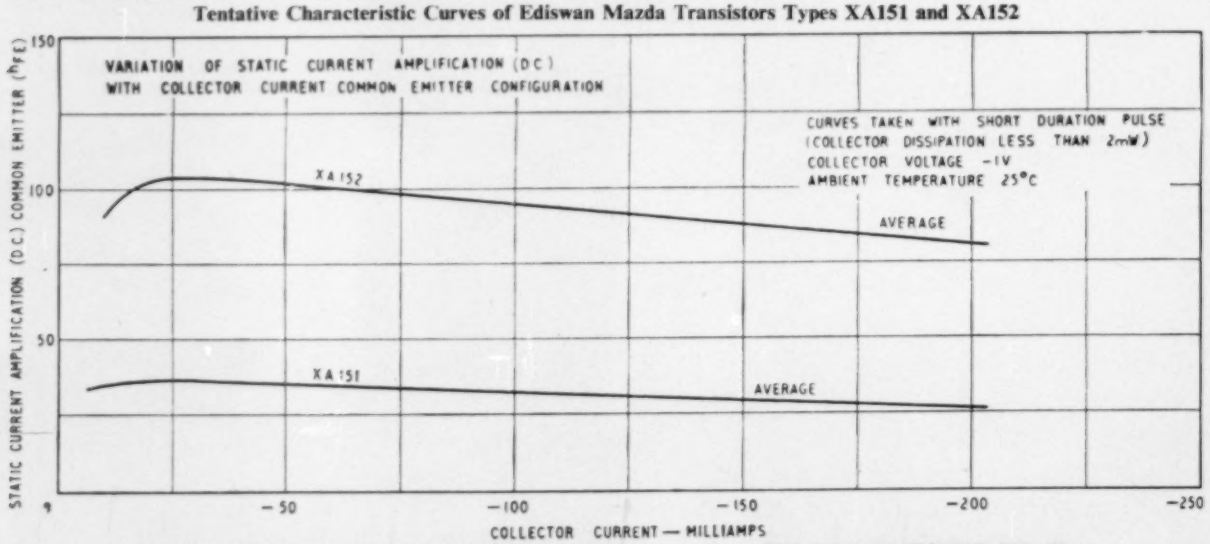
XA152 SWITCHING CHARACTERISTICS

	Typical Production Spread Min.	Av.	Max.
Static Current Amplification (d.c.)			
(V_{ce} = −0·25 V, I_c = −10 mA)	40	—	—
(V_{ce} = −0·25 V, I_c = −50 mA)	40	—	—
(V_{ce} = −0·25 V, I_c = −125 mA)	20	—	—
Collector/Emitter Saturation Voltage (volts)			
(I_c = −10 mA, I_b = −0·3 mA)	−0·04	—	−0·14
(I_c = −50 mA, I_b = −1·5 mA)	−0·06	—	−0·2
(I_c = −125 mA, I_b = −7·5 mA)	−0·06	—	−0·28
Base Resistance r_{bb}' (Ω)			
(V_{ce} = −6 V, I_e = −1 mA)	—	75	—
Cut-off Frequency f_1 * (Mc/s)			
(V_{ce} = −2 V, I_e = −10 mA)	5·5	—	—
Collector Depletion Capacity (pF)			
(V_{cb} = −6 V, I_e = 0)	—	11	16
Collector Time Factor τco (mμs)			
(V_{ce} = −0·25 to −4·25 V, I_c = −10·85 to −0·85 mA)	—	25	40
Saturation Time Factor τs (μs)			
(I_c = −10·85 mA, I_b = −0·23 mA)	—	0·9	—
βs† (I_c = −10·85 mA, I_b = −0·23 mA)	—	36	—

* Frequency at which the modulus of the common emitter current amplification is equal to unity.

† βs is defined as the collector current almost immediately available on closing the collector circuit, per unit steady base current.

Tentative Characteristic Curves of Ediswan Mazda Transistors Types XA151 and XA152



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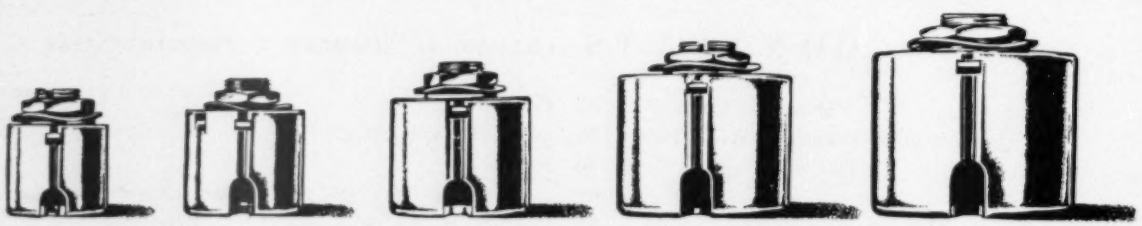

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Valves and Transistors

CORRESPONDENCE about the basic mode of operation of the junction transistor continues and this month we publish more letters on the subject. Most people seem to agree that it is not possible, by considering only the parameters of a transistor and its external circuit, to say that any particular form of connection is any more fundamental than any other.

If there is a fundamental mode, therefore, it would seem that it must be sought by considering the internal physics of the device. Anyone who attempts this would perhaps be well advised first to consider the triode valve, for the way in which it works is so much simpler. This may seem strange advice, for the common-cathode connection is so generally accepted as the basic one that it is hard to envisage that it might not be.

It must be remembered, however, that this could be no more than an historical accident. Early valves were all of the filament type and the earthed-cathode connection was thus the only one which permitted a single filament-heating battery to be used for several valves. This purely practical matter alone enforced the general usage of the common-cathode connection.

The question of whether or not there is one mode of operation which can be regarded as the fundamental is not highly important, in our opinion. What we do regard as important is that there should be one particular mode which is generally considered the normal one, even if it is an arbitrary mode selected as a matter of convention.

We have no doubt at all that this mode should be the common-emitter one. Just as with the valve, it is the most commonly used in amplifiers and, therefore, the one for which the small-signal parameters are most often wanted. The use of the same form of connection for both valve and transistor has also great advantages in teaching, for a unified theory can be built up. One other thing is needed to achieve this; mutual conductance should replace current amplification factor as a transistor 'constant'.

It is not always realized that the positive-grid triode can be represented by the same equivalent circuit as a transistor. It has an input resistance and, as grid current is in some degree dependent on anode voltage, it has a voltage feedback factor. It can even be regarded as having a current amplification factor! The ordinary equivalent circuit for a negative-grid valve is merely a special simple case of this more general one.

The only requirements for the use of the same equivalent circuit for valve and transistor are to adopt the same form of connection for both and to drop current amplification factor in favour of mutual conductance for the transistor. This last is, in fact, already being done by some workers. Taking a long-term view, we feel that the advantages in simplicity are considerable.

Experimental Travelling-Wave Tubes

C.W. MODELS FOR X-BAND WITH WATER-COOLED HELICES

By G. M. Clarke M.A., Ph.D.*

SUMMARY. *By using high-pressure water cooling of helices wound from tube, the advantage of high coupling impedance over a very broad band which the helix offers is retained, and the thermal dissipation proves to be sufficient for construction of c.w. broad-band amplifiers at X-band up to power limits set by other factors, such as perveance and backward-wave oscillation. Tubes are described with 150 W and 300 W c.w. output with gains greater than 20 dB over the band 7.5 to 11.5 kMc/s which, because of their small size, probably involve greater difficulties than would occur in higher power tubes using the same technique.*

A previous paper[1] has explored the limitations in c.w. power-handling capacity which are set by radiation and conduction cooling of helices, and, using these techniques, the boundary had been closely approached by various workers at the commencement of the work to be described here. For example, Siekanowicz and Sterzer[2] described a helix tube for 100-W output at S-band, limited to 0·1 r.f. duty-cycle, and although it was suggested that further development would enable continuous duty to be achieved, it was clear that to reach similar power output at X-band a new approach would be needed.

An investigation was thus commenced of the possibility of winding small helices from tube and employing some form of fluid cooling (either liquid or gas). It turns out that pure water is probably the best liquid (although at the expense of some decrease in thermal capacity there are silicone derivates which can run at high temperatures) and that steam is the best gas, on account of the high specific heat. It has been found that formation of steam in a water-cooled system has no drawbacks unless excessive life is required, and so the properties of both can be combined at the same time with an important additional contribution from the latent heat. The previous paper[1] outlined by means of scaling laws the range of applicability of this technique showing that, with the pressure system developed, there was no obstacle to its application in the upward direction to powers as high as could be developed within the limitations of voltage set by helix properties, and perveance set by gun design.

Difficulties are met rather in the low-power range, where the helix is small and the tubing has very thin walls. These were not fully appreciated when the tubes to be described were planned and they presented greater constructional difficulties than are likely to be present in larger power tubes using the same technique.

An alternative approach has been pursued elsewhere[3] based on the paralleling of structures and beams, and this has resulted in about 100 W at X-band[4] in a somewhat complicated tube, and this power is quoted as about the limit at X-band. This approach may, however, form a possible solution to the 'driver gap' problem which will exist with water-cooled tubes at frequencies somewhere above X-band.

Electronic Design

For the first trials, a cold circuit velocity of $c/8$ was chosen (4 kV) so that, at the operating voltage and with a gun microperveance of unity, something over 1 kW would be available in the beam. The dielectric loading factor of the support chosen was about 0·9 and it reduces dispersion somewhat. The helix dimensions can then be calculated, giving the helix pitch $p = 0{\cdot}037$ in. and mean diameter $2a = 0{\cdot}093$ in. A compromise then has to be made between the desirability for a large helix tubing diameter for maximum fluid flow and the consequent loss of impedance and restriction of the beam path. The effect on impedance of finite-sized nearly-round wires was calculated[5] using the developed approach and a pair of conjugate complex potential functions which allow arbitrary phase change per turn. The results of this calculation are shown in Fig. 1, where the impedance is plotted against the gap width to pitch ratio, for various values of the helix mean radius normalized to the free-space wavelength ($ka = 2\pi a/\lambda$).

A value of $ka = 0{\cdot}25$ was used in this valve and with the o.d. of the helix tubing at 0·018 in. we have $\delta/p = 0{\cdot}5$. The impedance is then 50 Ω as compared with a maximum of 66 Ω which could be obtained with smaller diameter tubing.

The values of Pierce's parameters[6] for a beam diameter equal to half the mean helix diameter are then: gain parameter $C = 0{\cdot}058$ and space charge parameter $QC = 0{\cdot}27$. The r.f. loss of this helix with copper plating was obtained by experiment and gives Pierce's loss parameter as $d = 0{\cdot}07$.

* *Ferranti Ltd., Edinburgh*

Watercooling

The simple transport equation

$$W = (4{\cdot}2)\, s\, p V \theta \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where W is the power removed (watts), s and p are the specific heat and density of the fluid, V is the flow rate in cc/sec. and θ is the temperature difference in °C, can be written for water.

$$W = 200\left(\frac{\theta}{48}\right) . V \quad .. \quad .. \quad .. \quad .. \quad (2)$$

from which it can be seen that if θ is 48 °C a flow of $V = \frac{1}{2}$ cc/sec. is sufficient to remove 100 W from the helix. $\theta = 48$ °C allows for a significant inlet temperature as would be required for a small heat exchanger and still allows a margin before the water boils.

The temperature difference is the average corresponding to complete mixing of the fluid and is the same whatever the power distribution applied to the pipe. However, the finite thermal conductivity of the fluid requires a radial temperature gradient and this will become appreciable at some power *density*, as distinct from the total power input. Some quite detailed analysis of heat transfer to fluids in laminar flow has been carried out[7,8] but for an estimate of the magnitudes involved we will use a simple analysis which assumes the radial and longitudinal temperature gradients to independent. It can then be shown that the temperature at radius r in a cylindrical pipe of radius a is given by

$$\theta_r - \theta_{axis} = \frac{W'r^2}{2\pi k' a^2} \cdot \left[1 - \tfrac{1}{4}\left(\frac{r}{a}\right)^2\right] \quad .. \quad .. \quad (3)$$

where k' is the thermal conductivity of the fluid and W' is the power applied per unit pipe length. Simultaneously, the average temperature across the pipe rises according to Equ. (1). A suitable integration of Equ. (3), weighted to allow for the velocity distribution across the section, then shows that the wall temperature θ_0 exceeds the average temperature by $11\ W'/48\pi k'$. Thus

$$\theta_0 = \theta + \frac{11\ W'}{48\ \pi\ k'} \quad .. \quad .. \quad .. \quad .. \quad (4)$$

Taking $k' = 4{\cdot}2 \times 0{\cdot}0014$ for water, the wall temperature will reach 48° C (the figure taken for the average temperature) above the average temperature when

$$W' = 4 \text{ W/cm.} \quad .. \quad .. \quad .. \quad (5)$$

With an inlet temperature of 30° C, the system would certainly be producing steam (126° C) if this power density occurred at the low-pressure end of the pipe though, if arranged to be at the high-pressure end, a few hundred p.s.i. would suppress this. Little advantage can be taken of pressure suppression of steam as the water must always reappear at the low pressure before it reaches the heat exchanger.

With a helix velocity of $c/8$, there is approximately eight times as much tube length compressed into the axial length of the helix (z direction) and so multiplying Equ. (5) accordingly the power density is

$$W'' = 32 \text{ W/cm. axial length} \quad .. \quad .. \quad .. \quad (6)$$

This can be related to the r.f. output power P_o and attenuation α db/cm via

$$W'' = dP/dz = \left(\frac{\alpha}{8{\cdot}68}\right) P_o \quad .. \quad .. \quad .. \quad (7)$$

Taking $\alpha = 1$ dB/cm, $P_o = 100$ W

$$W'' = 11{\cdot}5 \text{ W/cm.} \quad .. \quad .. \quad (8)$$

which according to Equ. (6) is three times less than would produce the radial temperature rise used in deriving Equ. (5). Alternatively, 3% of a kilowatt beam could be allowed to arrive in 1-cm length of the helix giving $W'' = 30$ W/cm. The calculations outlined above show that in the system contemplated with about $\frac{1}{2}$ cc/sec. flow, the condition most likely to drive the wall temperature high enough for steam formation is the arrival of a few per cent of beam interception in less than 1 cm length of the helix, while the power density produced by attenuation dissipation is some 30% lower. As indicated later, it was found possible to run considerably higher in power than that giving first production of steam, a great advantage then being obtained from the latent heat of boiling.

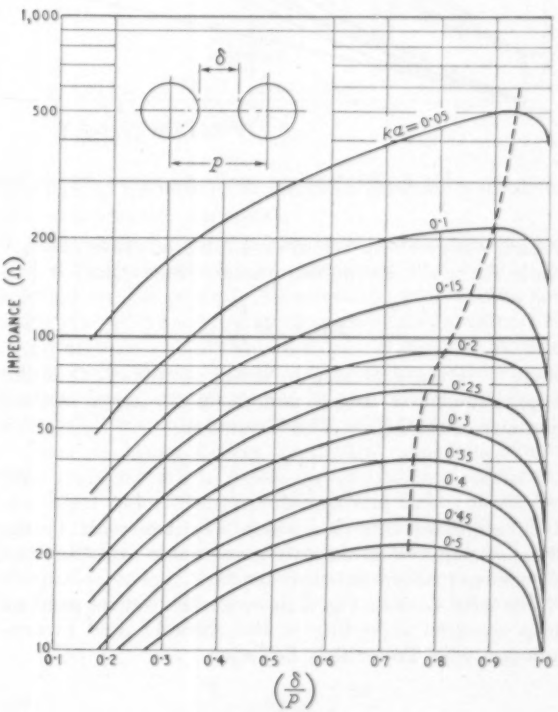

Fig. 1. *The impedance in the plane of a developed helix with nearly-round wires*

Pressure Required

Early experiments immediately showed that the pressure drop in a helix is considerably greater than calculated from the classical flow in the equivalent length of straight tube, even at low flows where the flow was certainly streamline. The literature reveals that this effect has been known for a considerable time and an empirical formula for the factor C (not to be confused with Pierce's C) by which the flow impedance is in-

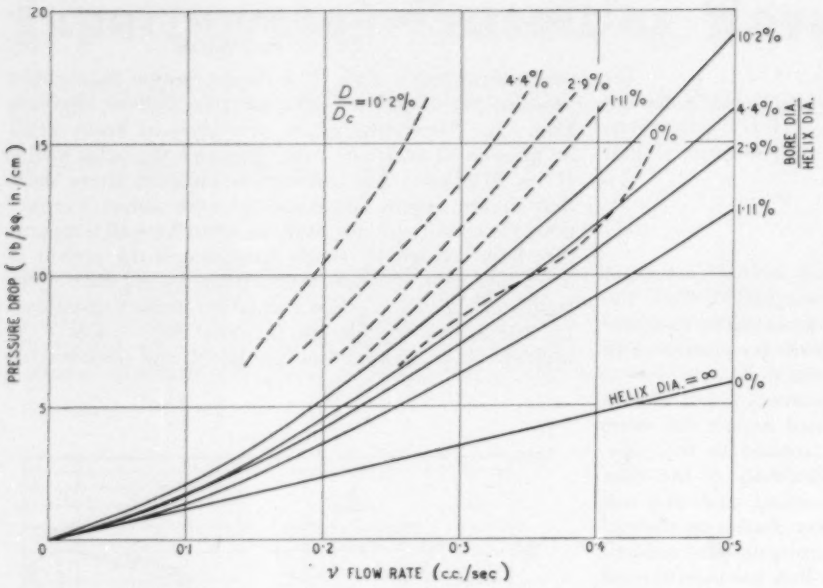

Fig. 2. *Theoretical and experimental pressure drops in helical flow*

creased was given by White[9] as a function of Re $(D/D_c)^{\frac{1}{2}}$ where Re is the Reynolds number. For water

$$\mathrm{Re} = 50\,\frac{V}{D}$$

where V is the flow rate in cc/sec as before, D is the pipe diameter in inches and D_c is the mean coiling diameter of the helix. The formula is

$$C = \left[1 - \left\{1 - \left(\frac{11{\cdot}6}{\mathrm{Re}(D/D_c)^{\frac{1}{2}}}\right)^{0{\cdot}45}\right\}^{2{\cdot}22}\right]^{-1} \quad .. \quad (9)$$

As C increases with Re which increases with V, the pressure required is not only greater due to coiling but is no longer proportional to the flow.

The solid lines in Fig. 2 show the theoretical pressure drop required according to this formula used in conjunction with Poiseuille's Equation

$$\frac{dP}{dz} = 14{\cdot}19\,\frac{V}{(100D)^4} \quad .. \quad .. \quad (10)$$

for $D = 0{\cdot}0105''$, and four different coiling diameters D_c including infinity.

The smallest coiling diameter shown corresponds to that used in the travelling-wave tube and shows a pressure increase of some three times for 0·5 cc/sec. flow.

Experimental measurements taken on this range of helices shown dotted in Fig. 2 are displaced towards still higher pressures. In comparison with the straight pipe, the effect of coiling is reasonably described by the calculated factor C, but the straight pipe itself has a greater pressure drop than expected. The reason for this is not known but, as the pressure depends on the fourth power of the diameter, the discrepancy could be accounted for by an error in the diameter, or possibly the unknown effect of surface roughness which is appreciable in such small-bore tube.

As the helices are some 50 cm. stretched length in the tube, it can be seen that pressures exceeding 1,000 p.s.i. are required to obtain the required flow. The mechanism which gives rise to the increase in flow impedance with coiling is the appearance of a secondary flow superimposed on the axial flow in the pipe[10] as depicted in Fig. 3. This is set up by the centrifugal force acting away from the axis and also has the effect of stabilizing the flow against turbulence quite markedly[11]. Furthermore, it is clear that such transverse motion will increase the heat transfer from the wall to the fluid above that calculated with Equ. (3).

Pump Design

No suitable pumps were found to exist for the service required here and one was specially developed. It is a double-acting diaphragm pump with ball-valve gear and is shown in Fig. 4. A central motor-driven cam displaces alternately two steel pistons in contact with synthetic rubber diaphragms which form the base of the two conical pressure chambers. The pistons are faced with p.t.f.e. to reduce to a minimum the abrasive action on the diaphragm. Communicating with the centre of the pressure chambers are the blocks (shown to the right) which contain the valves. On the inlet stroke water is admitted through the inlet valves from a low-pressure reservoir. The pressure in this reservoir

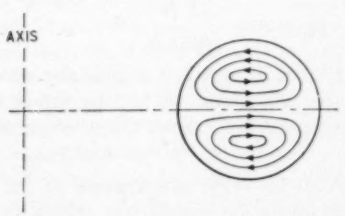

Fig. 3. *Secondary transverse flow in a helical pipe*

is applied during the filling operation and it is then sealed off, and remains at this pressure throughout the life of the pump. To this end, the pumps require to be leak-searched to vacuum standards before filling. The low pressure in the reservoir (approx. 50 p.s.i.) prevents cavitation on the inlet strokes and assists in maintaining proper seating of the valves.

On the pressure stroke, fluid is discharged into the high-pressure reservoir, whose pressure is maintained by a heavy spring in the upper housing which operates on a further piston in contact with the reservoir diaphragm. From here the fluid is fed to the helix and on exit it is returned to the low-pressure reservoir which is provided with a heat exchanger. Although gauges

have been used on both high- and low-pressure sides on occasion, the high pressure can be monitored from the position of the reservoir piston by means of the rod shown, which is used to operate a safety device to ensure that the tube power supply cannot be operated without helix cooling water.

Apart from the diaphragms, all internal parts of the pump are of stainless steel to minimize corrosion, and high-purity water is used for the filling. Bleed points are supplied at the top of each pressure reservoir, for the pump will not operate if any air is trapped in the high-pressure side during filling.

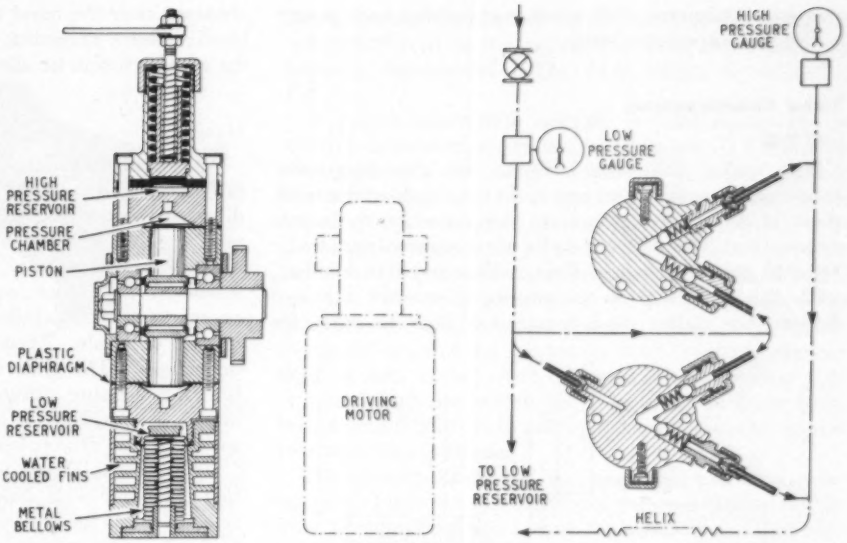

Fig. 4. *Cooling-pump section*

The energy absorbed in the flow is small (of the order of a few watts), and the motor drive is called upon mainly to supply torque impulses up to 10 lb in. twice per second, and it takes a continuous input of 240 V 150 mA, most of which appears as heat in the motor casing. Early pumps with neoprene diaphragms ran for only a few hundred hours before the diaphragm wore away, but current models with 'Duthane' diaphragms have been taken down for inspection at 1,000 and 2,000 hours and show insignificant signs of wear.

An external view of the pump mounted on a 12-in. square base plate is shown in Fig. 5.

Simulated Thermal Performance

In order to simulate as far as possible the conditions which may arise in the operating travelling-wave tube without the expense of testing complete tubes to destruction, experiments were performed in air, with a few turns at the end of a test helix fed with mains-frequency current. The helix surface temperature is difficult to measure because of the small size and, although temperature sensitive paints were used, the gradient towards the current leads made this unreliable. Most tests were thus performed by running up to the first appearance of red heat, and using this as a setting-level for comparison between different helix materials. A dissipation level of 400–500 W was reached at this temperature regardless of the helix material.

After trials over long periods running near the maximum power limit, it was found that stainless steel was inferior to nickel copper alloys, although the attack of the material was less. Stainless-steel tubes invariably failed eventually by blocking of the bore, while cupro-nickel helices after approximately twice the life, punctured. In the latter case, the fluid was found to be considerably contaminated with copper salts and, presumably, the difference in the two cases is that with stainless steel the corrosion products are insoluble and eventually clog the bore, while with cupro-nickel they circulate in solution until sufficient material is lost from the wall to cause puncture.

Unfortunately, cupro-nickel helices are extremely soft and difficult to handle in production and had to be rejected on this ground. However the power-limit tests referred to above were at levels considerably greater than could arise in a 1-kW beam tube; at the 100 W level (which still allows a good safety factor with a well-focused beam) stainless-steel helices would run for at least 2,000 hours and so could be adopted.

Tests on larger helices for 10-kV operation wound from 0·030 in. o.d. by 0·020 in. i.d. stainless steel have also been run. The same pump was used run at five times the speed (140-cc/min displacement) with a larger torque motor and continuous removal of 1-kW dissipation in four turns was obtained without difficulty for long periods. The power limit could not be reached

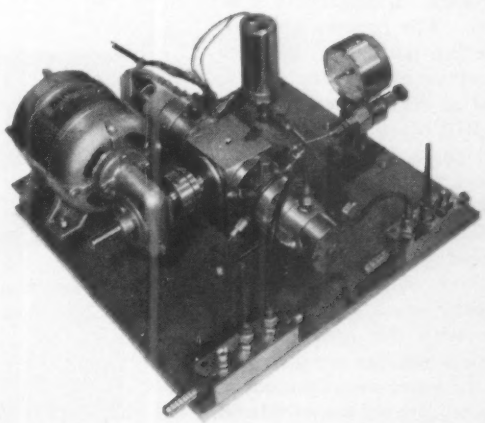

Fig. 5. *General view of cooling pump*

owing to equipment difficulties in supplying such power at the low-impedance level.

Tube Construction

Couplings

The major difference between this travelling-wave tube and a conventional one is that the ends of the helix must be available external to the valve for the water connection. The coupled helix arrangement had to be rejected on two grounds. First, calculation [12] shows that, with the dielectric of an envelope between the two helices, the radius ratio required is near to unity for

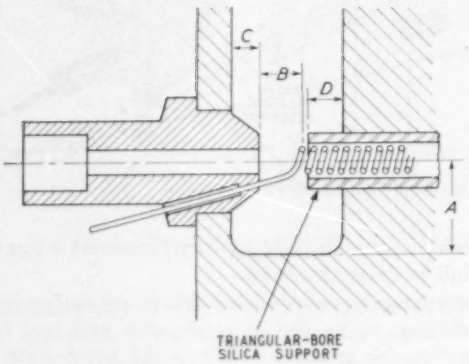

Fig. 6. *Couplings*

broad-band operation, and with the necessity for tubing rather than tape to be used, the envelope becomes very thin and fragile at the coupler. (If the envelope is not between the two helices the important advantages of adjustability and fabrication which the coupled helix arrangement offers are lost.) Secondly, at least at the output end the coupling helix also needs to be water-cooled, which is an embarrassing complication.

The solution finally adopted was to use evacuated waveguides so that there is no dielectric material in the high-power density region of the junction, and to use a 'door-knob' arrangement as in Fig. 6. The optimum values of the four parameters labelled were obtained by study with a scanning reflectometer. The most critical parameter is B, the aerial length, and the bend into the first turn of the helix must be made on a jig to obtain reproducibility. A variety of different velocity helices from 4 kV to 10 kV have been matched into a No. 10 waveguide with this arrangement. The best matches are obtained with the lower-velocity models, being no greater than 1·3 from 7·5 to 11·5 kMc/s. With the 10-kV helix, although even better performance can be obtained over the band 8–10 kMc/s, to obtain the wider band without exceeding 2:1 at the ends the v.s.w.r. in the middle has to be allowed to rise to 1·5.

Windows

To maintain the large bandwidth required, a virtually full-aperture window is necessary with the minimum of dielectric material. Originally, thin-mica windows were used sealed into chrome-iron frames with a low-melting-point glass. Electrically, these performed very well, being less than 1·05 s.w.r. over the band, but trouble was experienced with windows developing leaks during the baking schedule. Thus ultra-thin glass windows were developed. The glass thickness was reduced until the bursting pressure difference was about 40 p.s.i. which allows the application of forced air cooling if required, as well as the atmospheric-pressure difference.

The glass thickness was then 0·010 in., the material being Kodial with Nilo frames. Although the s.w.r. reaches 1·1 at the ends of the band, this can be tolerated, and the windows have been very satisfactory. Unblown, they will pass 500 W c.w. and, blown, 2 kW c.w. has been passed without failure. They will also handle 1-MW pulses at 1-kW mean.

Attenuator

For a power tube, the attenuation region should be as short as possible and, to cope with the eventuality of significant reflected output power, must be thermally rugged. To meet these requirements, a number of lossy ceramic materials were investigated, including porous materials impregnated with carbon. Many of these materials are extremely hard and difficult to work, but it was eventually found that heavy-duty ceramic resistor material which is satisfactory in vacuum could be obtained in a large range of volume resistivities. This material is made by compression of carbon and ceramic powder followed by sintering, and the resulting product is machinable with tungsten-carbide tools. Theoretical work[13] proved to be of limited use in estimating the volume resistivity required, as experiments showed a large change of attenuation with change of contact with

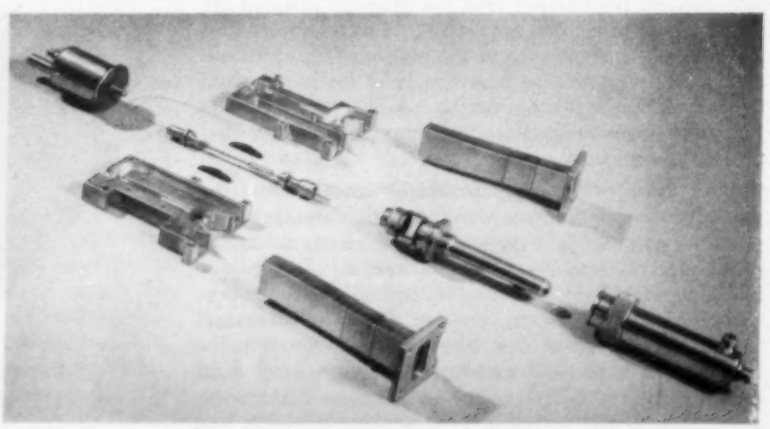

Fig. 7. *Water-cooled travelling-wave tube before assembly*

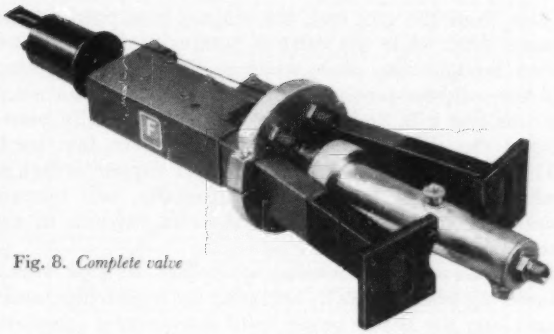

Fig. 8. *Complete valve*

the helix turns. A number of different tapers were studied experimentally to obtain a satisfactory match.

All have a parallel portion near the middle which must be in intimate contact with the helix. When this is so, the attenuation exceeds 50 dB in a 3-mm length. If this contact is not positive the loss drops sharply to 15 dB, which is inadequate. The necessary contact cannot be achieved reliably if the attenuator is a solid tube and slides over the helix, unless the fit is so good that there is danger of disturbing the accurately-controlled helix pitch. Thus the attenuators are made in two halves. They are unaffected by heating to over 1,000 °C in brazing operations on the assembly.

Body

The body is constructed by milling two halves from a sheet as shown in Fig. 7. The helix assembly including the door-knobs and triangular-section silica supports is then mounted between the two halves and, with temporary clamping, the match is examined and any trimming necessary done on the aerial length. The helix assembly is then removed and the door-knobs brazed to the helix. This is the most difficult part of the assembly as, without careful control of the conditions, either a satisfactory vacuum-tight seal is not obtained, or brazing material will permeate the thin-walled tubing rendering it useless.

The attenuators are fitted into the two halves which are brazed together about the helix assembly together with copper waveguide legs which have been previously furnace-brazed to Nilo sections at the window end. This is so that eddy-current brazing can be applied to seal on the window assemblies which cannot be furnace brazed. A complete valve is shown in Fig. 8.

Gun and Focusing

The original gun was designed according to Mueller[14] and is for unit microperveance. It was built and tested in glass envelopes and found to be satisfactory apart from rather a small clearance between condensing electrode and anode face, and this was increased with little change in electrostatic performance. The gun was then transferred to metal envelope, mainly in order to enable an external polepiece for Brillouin focusing to be used. This required positioning as close as possible to the anode face and would have been difficult in glass. At a later date, when the focusing performance had been thoroughly assessed on simulators, this polepiece was transferred to the anode face itself, thus removing focusing adjustment. The final design is shown in Fig. 9.

A barium-aluminate cathode is used owing to the current-density requirement, which is above 1 A/sq.cm. The gun area convergence from cathode to nominal beam diameter is 25 : 1 though, in conjunction with the focusing fields applied, it would appear from tube performance that it is in fact greater than this. The Brillouin field for the nominal beam diameter is 1,200 gauss. First tests gave poor transmission even with fields exceeding 2,000 gauss and this was found to be due to the influence of the a.c. heater field. With d.c.-heater field it was found that improved transmission was obtained with the heater field opposing the main field, but sufficient field could not be obtained from this source to obtain the optimum.

The envelope material was thus changed to a non-magnetic cupro-nickel alloy and external fields applied from an auxiliary coil round the gun. Increasing this field drives the optimum main field to higher values with continuous improvement of the transmission. Without auxiliary field the optimum main field is about 1,600 gauss and 95% transmission is obtained, but the heater must run from d.c. As the auxiliary field is increased the main field may be increased until more than 99% transmission is obtained with the main field

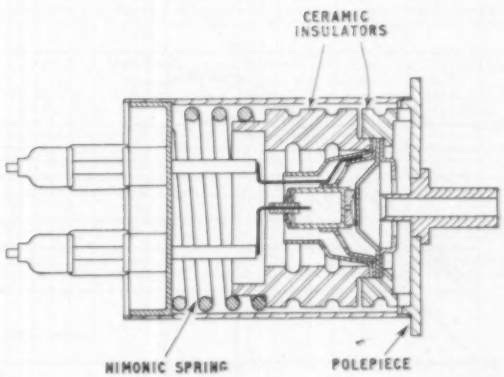

Fig. 9. *Unit microperveance gun*

at 2,000 gauss, the heater field then being unimportant. It is unnecessary to use such good transmission in the presence of the water cooling, and one can thus avoid the high solenoid dissipation (1·2 kW) which this entails. Operation at 1,800 gauss 98% transmission with 800 W in the solenoid has been the rule in life tests though lower transmission than this with further economy in solenoid power has been used. A shorter solenoid might be used if an exit polepiece were adopted.

Processing

Difficulties with gas evolution from the collector on first running the beam necessitate operation of the valve while still on the pump. As the valve is pumped through the gun and the magnet can only be passed over the

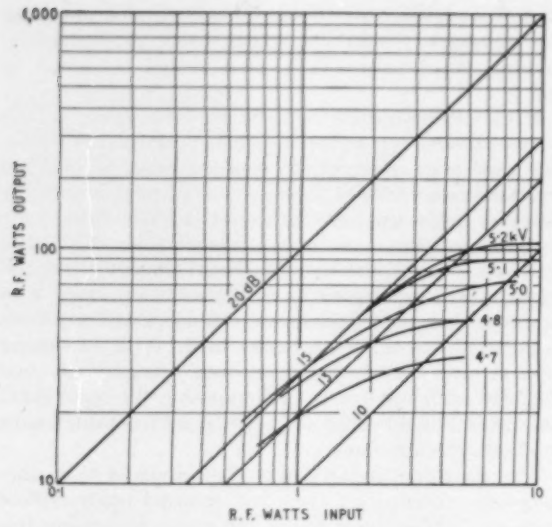

Fig. 10. *Power output of Mk. I tube*

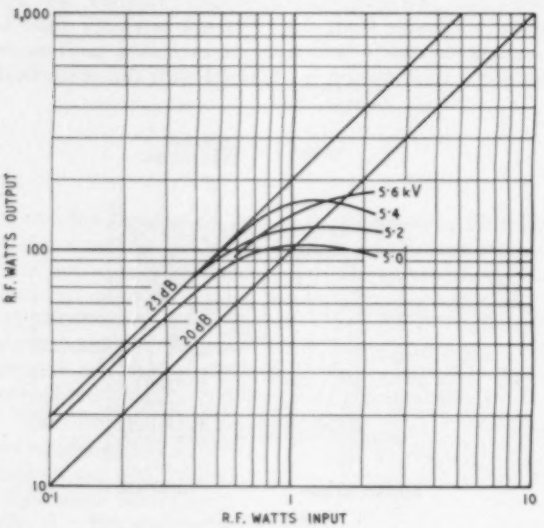

Fig. 11. *Power output of Mk. II tube*

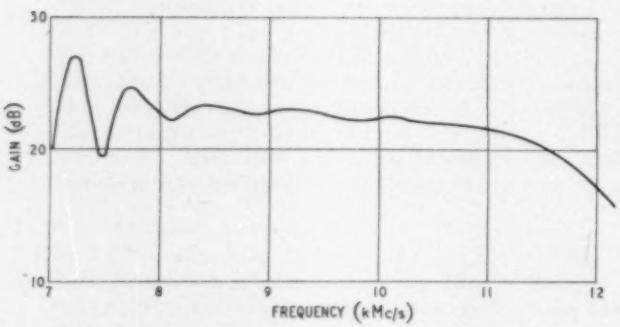

Fig. 12. *Small-signal gain of Mk. II tube*

valve from the gun end, the magnet is trapped on the pump stem while the valve is pumped and baked, and then brought into place when required. Ion focusing in the collector prevented expansion of the beam when it was first run, and the excessive power density punctured the thin-wall collectors which were first used. These were replaced by thick-walled copper collectors which although locally melted initially, will operate until the valve is hard and the beam expands in the collector.

An automatic processing controller is used which limits the pressure rise by adjusting the oven temperature and later the heater power until activation is complete, the whole schedule taking 12 hr with 460° baking temperature.

Performance

Mk. I

First valves, built according to the dimensions of the earlier section (Electronic Design), showed somewhat lower gain than expected, falling with frequency across the band accessible to the swept driving oscillator (7·5 to 11·5 kMc/s). At the low end the small signal gain was 17 dB falling to 13 dB at the high end. The optimum helix voltage was higher than the 4·5 kV expected, being 4·9 kV.

This behaviour could mostly be accounted for by too small a beam filling for the somewhat large radial propagation parameter γ_a of 2.0. Certainly, on these first valves, the absolute maximum in focus performance was used to minimize risk on the untried cooling system. However, when driven at 8·7 kMc/s where the small signal gain was 15 dB, outputs above 100 W could be obtained thus realizing the target. The optimum voltage for maximum saturated gain was rather flat and occurred at 5·2 kV with 300 mA cathode current, the gain being 13 dB. This performance is summarized in Fig. 10.

Mk. II Tube

In this tube the helix dimensions were modified to the following: pitch = 0·0296 in. and mean diameter = 0·089 in. representing some 20% reduction in pitch and 15% in diameter. The active helix length was unchanged. The pitch change increases the gain via the increased number of wavelengths in the helix and the diameter change increases the beam filling. The calculated change in operating voltages is quite small, the cold velocity reduction being offset by the rise in coupling and thus *QC*. A satisfactory increase of gain was obtained and although similar power outputs appeared at the voltages which were optimum in the Mk. 1 tube, with this tube the operating voltage can be raised to 5·4 kV (350 mA) giving over 150-W output with 21-dB gain. Power exceeding 170 W can be obtained with still higher voltages but only at the expense of reduced gain at low drives. The performance is summarized in Figs. 11 and 12.

Mark III Tube

Having established the capability of the water-cooling system in sealed-off tubes (the beam was defocused in

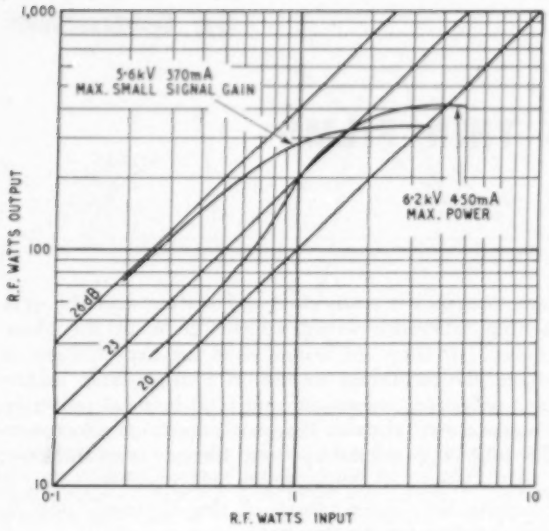

Fig. 13. *Power output of Mk. III tube*

some tubes as far as 200-W interception without harm) it was decided to attempt to apply more beam power in the same body and gun by increasing the helix pitch to 0·033 in. leaving all other dimensions unchanged.

The power output of this tube is shown in Fig. 13. The optimum helix voltage for maximum small signal gain is 5·6 kV and over 300-W output was obtained with just over 1-W drive. 400-W output was reached at 3-W drive with a higher beam voltage (6·2 kV) but this was not then optimum for small signal gain which dropped to just over 20 dB.

A general view of a valve set up in its magnet with cooling pump is shown in Fig. 14.

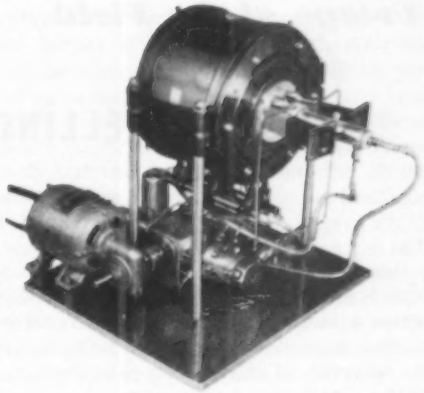

Fig. 14. *General view of water-cooled helix travelling-wave tube with solenoid and pump*

Acknowledgement

Key contributions to this project were made by G. B. Lees who designed the cooling pumps, and by R. W. Murray, J. Fisher and W. Duffy who constructed the tubes. The author is indebted to Ferranti Ltd., for permission to publish this paper.

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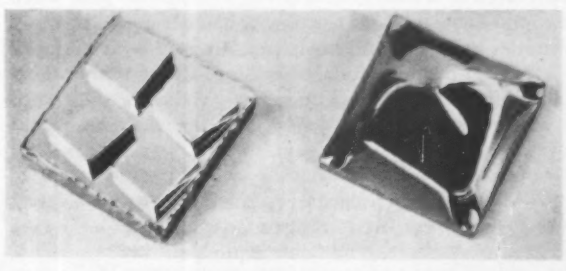
SILICON TRANSISTOR MANUFACTURE

A process for the construction of silicon transistors is being developed by the Research Laboratories of the General Electric Co. Ltd. This organization is, incidentally, now called the Hirst Research Centre, after its founder forty years ago. The new process is called the re-grown diffused technique and it is expected that it will make the production of silicon transistors as economical as that of germainium types.

Two blocks of silicon, one heavily doped with arsenic (n-type) and the other with boron (p-type), are melted and fused into a larger block of silicon lightly doped with arsenic. When the temperature is lowered, the molten highly-doped material re-crystallizes on to the unmelted region to form once again a single crystal. In this re-grown region, the arsenic predominates and the silicon is consequently n-type.

The material is then baked at a high temperature and, during this baking, boron diffuses out of the re-grown region to form a p-type base region with a thickness that can be controlled to 0·0001 in. or less. The process requires only simple apparatus and the characteristics of the final device can be accurately controlled.

At the end of the process, the material is of the n-p-n structure and of a pyramid shape. It is then cut into bars of the dimensions required for the transistors. Early transistors produced by this process have cut-off frequencies in the range 20–30 Mc/s and the characteristics are stable even when operated for long periods at 200 °C.



Blocks of heavily doped n- and p-type silicon are shown on the left resting on a block of lightly doped n-type silicon. On the right can be seen the pyramid-type structure obtained after the fusion processes

TUNNELLING AND THIN FILMS

Sometimes, when I open a book on theoretical physics, I experience a kind of sinking feeling. It all looks so abstruse, and a good deal of it actually is. The real difficulty, however, is in realizing that the writers are trying to be intelligible; that when they speak of a wave, they *mean* something that is to be interpreted as having the ordinary properties associated with the simplest kind of wave, and when they speak of a particle they *mean* something that can be dealt with by the particle mechanics of Newtonian or of relativity physics. Sometimes it looks as if they are trying to have the best of both worlds, as when they start working in terms of momentum and Hamiltonians, or operate Fourier transforms; but here they are really farming out the hard work to a mathematician or a computer and, when it comes to sorting out the result, it is always in terms of particles and potential fields that interact with them, or of waves and the boundary conditions that enable them to survive. Usually both together. Which of the results is selected for attention depends on the further use that is to be made of it.

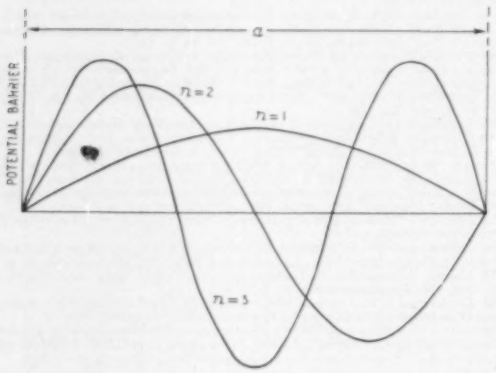


Fig. 1. *The first three stationary-waves for an electron in a 'box' bounded by a very high potential barrier on each side. If a is the length of the box, the only possible wavelengths λ are 2a/1, 2a/2, 2a/3 . . . . 2a/n. These fix the corresponding velocities, and hence the energies of the quantum states n = 1, n = 2, n = 3 . . . . etc. The diagram has a double ordinate-scale, in that the vertical line indicates the potential barrier, which is not related to the arbitrarily drawn wave amplitude*

One often wonders whether, when people speak of waves, they are thinking of electromagnetic waves (which are very well understood), de Broglie waves which are the mechanical equivalent of particles, Schroedinger probability waves which help to locate the regions inhabited by particles, or something that satisfies a possible solution of a wave-function equation. For many purposes it really does not matter, and the only essential difference would appear to be in the phase velocity. If they are waves, then we expect them to behave like the things we see on a ripple-tank, undergoing reflection, refraction, and total internal reflection at boundaries in the way Huygens' construction forecasts. The following quotation from Bleaney and Bleaney, p. 515, seems to amalgamate all possible ways of regarding the 'wave-nature of the electron', and to justify settling for the simplest:

> "The fact that electrons and other particles have a wave-like aspect is well established from experiments on electron and neutron diffraction; the wavelength associated with a particle of linear momentum **p** is given by the de Broglie relation $2\pi/\mathbf{k} = \lambda = h/\mathbf{p}$, where **k** is the wave-vector which is parallel to the direction in which the wave is travelling, and h is Planck's constant. For example, the wavelength is $12{\cdot}3 \times 10^{-8}$ cm for an electron of kinetic energy equal to 1 eV. Free electrons confined within a metal rebound from the surface of the metal without losing any energy; on the wave aspect, this means that the waves are totally reflected at the boundaries, and standing waves are set up. These standing waves are the allowed solutions of the wave equation for a particle in a box, in the same way that standing waves of certain wavelengths are allowed in a waveguide resonator."

I think that three different concepts of wave itself are present in this quotation. But perhaps the most noteworthy thing in this quotation is the figure for an electron of energy 1 eV. The temperature for an electron to reach such an energy in thermal equilibrium is about 10^5 °K; it is clear that the electrons in a piece of metal at ordinary temperatures in thermal equilibrium with the ions of the crystal lattice of the metal will have energies of the order 10^{-4} eV or so. This means de Broglie wavelengths of the order of 10^{-3} cm, which leads to the question of what happens to the electron-waves in a film which is thinner than this? The basic answer, and one which is supported by measurements of the variation of specific heat with temperature (to which the electrons should make a significant contribution if they were in thermal equilibrium, but don't because they aren't) is that the majority of the electrons in a metal (or any other solid) at ordinary temperatures are effectively isolated thermally from the crystal lattice, and have much the same energies (which in the upper ranges are very large ones) as they would have at absolute zero. The upper limit for a given metal at absolute zero, the Fermi Level, is usually several eV; in a metal, electrons with Fermi-level energies are those

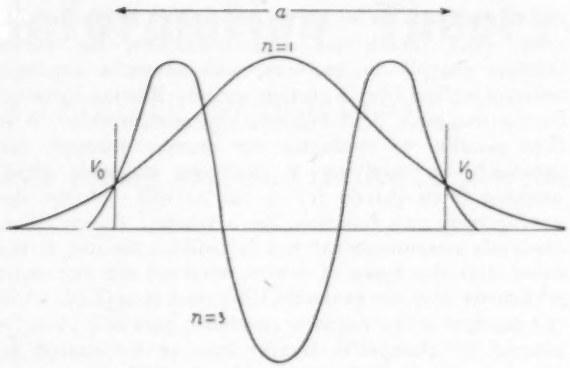


Fig. 2. *Two of the energy-states, or stationary-wave patterns, for a finite potential barrier V_0. The 'tails' outside the boundary of the box indicate the probability of finding an electron outside it, a tunnelling through the potential barrier, or merely the penetration incidental to total internal reflection, as you choose to phrase it. As in Fig. 1, the vertical scale is mixed; V_0 indicates the potential barrier, and the height of this is not related to the ordinate scale for the amplitude of the waves*

responsible for electrical conductivity. These can be perfectly happy in a thin film, in which less energetic ones could not survive.

Later (and, in spite of broken undertakings in the past, I am determined to do this in the next article for certain) I will go into the business of the Fermi Level, and the more intricate matter of the Fermi Surface. Also I shall have to discuss some of the difficulties that surround the establishment of a theory of superconductivity, which ties up with this. For they will be needed in order to discuss the significance of the thin-film tunnelling experiments that are to be described briefly in the present article.

Single Electron in a Box Bounded by a Potential Barrier

In other words, an electron on its own travelling to and fro between the walls of a piece of metal, where the potential barrier at each face reflects it back. Regarding the electron as a wave, of wavelength $\lambda = h/mv$ there are a number of stationary-wave patterns, each with the boundaries as nodes (Fig. 1), which determine possible values of v and hence of the kinetic energy. And on the simplest wave-picture, which assumes total internal reflection, we have to remember that we are only going to get this total reflection if we do not do anything that might abstract energy very close to either boundary; for (see, e.g., Bleaney and Bleaney, p. 257, or any book on physical optics) total internal reflection does involve a penetration of energy beyond the boundary, tailing off rapidly in an exponential manner within a few wavelengths, and its subsequent return in the right phase if it isn't interfered with. A good example of what I mean here is the frustrated-internal-reflection filter used in optics, where a very thin film of low-refractive index sandwiched between two high-refractive index blocks transmits light of a selected wavelength range that would have been totally reflected if the film had been thicker.

Going into things a little deeper, we can see an obvious flaw in the optical analogy to Fig. 1. For, if the potential barrier is very high indeed, then the wavelength of the disturbance outside it will be very short, and the excursion beyond the boundary very small. But (Fig. 2) if we suppose the boundary to be of height V_0, which is larger than likely values of the energy of an electron E but still not huge, then the Schroedinger *wave equation* shows that the *wave-function* has finite values beyond the boundaries. You can interpret Fig. 2 in one of two ways—either by considering that the tails of the curves beyond the boundary indicate the probability of finding the electron *there* (not well beyond it, but within the length of the tail) or else, on the more literal de Broglie wave picture, as total reflection that is frustratable if we hunt for electron-waves within the tail.

There is one other point that should be cleared up before going further. The first demonstration of this 'tunnelling' effect of particle-waves through a potential barrier was in the 1930s, when Gamow showed that this was a possible mechanism by which α-particles could escape from the nuclear potential field, and Cockcroft and Walton effectively worked this in reverse by disintegrating lithium nuclei using relatively low-energy bombarding protons. The α-particles don't hang about within a few wavelengths of the barrier, but shoot out with huge energies. This is a quite different case, however, because once through the barrier they get the full benefit of the Coulomb field of the rest of the nucleus.

Ivar Giaever's Thin Film Experiments

In November 1960, the General Electric Company of Schenectady issued a report of some recent work by Ivar Giaever, which is described as a discovery that yields new concepts, may make possible a new family of electronic devices, and is already adding to fundamental knowledge of both the tunnelling effect and superconductivity. It started as an exploration of the wider possibilities of the tunnelling effect (that is, passage of electrons through a potential barrier in the way indicated above for the surface of a metal) at a semiconductor p-n junction, which is the basis of the new semiconducting tunnel diode. The device consists of a thin film of insulating material sandwiched between two metal films, one or both of which may be in the superconducting state.

A thin film of aluminium (Fig. 3) was first evaporated on to a glass slide. A thin film of insulating oxide, about 10^{-7} cm thick, was deposited on this by exposure to the air. Next, a film of lead was evaporated over this, at

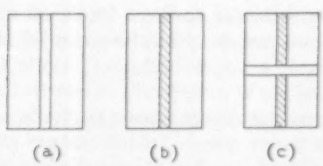


Fig. 3. *Stages in preparing the tunnel film; (a) a strip of aluminium is evaporated on a glass slide, (b) the strip is oxidized, and (c) the strip of lead is evaporated across the aluminium strip, sandwiching the oxide film between the two metal strips*

right angles to the first film, so that there was a thin film of insulator sandwiched between the intersection of the strips.

Superconductivity, a property shown by both lead and aluminium, and by a number of metals and their alloys but not by all metals, is characterized by the complete disappearance of ohmic resistance below a certain critical temperature T_c the value of which depends on the metal. No satisfactory theory of it seems so far to have been established, chiefly because it must involve some interaction between the conduction electrons and the lattice-vibration phonons, and this is the first thing that the solid-state experts have to neglect in order to make their sums manageable by the ordinary kind of computer. The recent theories of Bardeen and

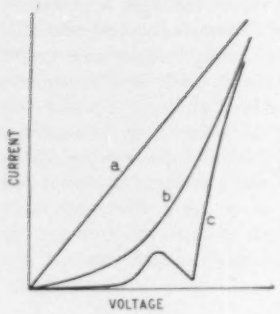


Fig. 4. *Current-voltage characteristics for tunnelling through a thin insulating film; (a) tunnelling between two normal metals, (b) tunnelling between one normal metal and a superconductor, and (c) tunnelling between two superconductors, showing negative resistance over a certain range*

of Fröhlich suggest that the phonons may co-operate with the electrons by a sort of resonance or synchrotron effect instead of scattering them, and this is the line along which ideas are running.

For lead, T_c is 7·2 °K, and for aluminium it is 1·2 °K. This is in the absence of a magnetic field, for a magnetic field lowers the critical temperature, so that applying or removing a strong magnetic field can be used to switch off and switch on the superconductive property below T_c at will, a feature of great technical importance.

A metal in the superconducting state is a perfect diamagnetic, for magnetic flux cannot penetrate within it (the Meissner effect); no thermoelectric effects are obtainable between superconductors; and there is no Thomson effect. In many ways it is just like a box containing free electrons, with nothing to keep them there except the potential barrier, the exclusion principle and Fermi-Dirac statistics. Thermodynamically, people seem very happy about its behaviour; but thermodynamics is the macroscopic tool, which is used when we do not know (or do not have to enquire) what is going on beneath the surface. And all that we know about these two metals is that between 7·2 °K and 1·2 °K one of them is a superconductor, while below 1.2 °K they both are.

Fig. 4 shows the current–voltage characteristic for the sandwich in three cases. In (a), above 10 °K, there is the normal Ohm's law behaviour, with nothing particularly odd about it except that this is a current through an *insulator*. In (b), with lead superconducting, there is a marked change, the small current flowing when the potential difference is small suggesting a reluctance by the superconductor to accept electrons. While in (c), when both metals are superconductors, the curve changes sharply in gradient, and shows a negative resistance effect over a certain region. Bearing in mind that a magnetic field destroys superconductivity, it is thus possible to modulate the current through the sandwich by applying a changing magnetic field, jumping from (b) or (c) to (a) at will, so that the arrangement can function like a triode. Curve (b) is obviously unsymmetrical, but (a) and (c) are not, in the sense that the types of curves obtained are the same whichever way the potential difference is applied. And the position of the negative resistance part of (c) can be altered by change in temperature or by change in magnetic field. This account is a faithful summary of the General Electric Company's description of the device. It now remains to try to interpret its action in simple terms, as far as I can; and it seems to me to be a sort of exercise in Angstroms. For the thickness of the insulating film, the separation of the ions in the metallic film lattices, and the wavelength $\lambda = h/mv$ of the electrons concerned are all of the same order of length, about a dozen Angstroms or so. We will start with the insulating film first.

Optical experiments on thin dielectric films like that of the insulator suggest that there is usually a lack of continuity, and that they are full of holes. The function of the insulator is thus possibly simply to keep the metal surfaces from contact with one another. And these are close enough for the tunnel-tails of Fig. 2 to extend from one to the other. The applied potential difference, instead of dissipating energy in the oxide film may, by lowering the V_0 potential barrier on the positive side, simply shift the projecting tail of the wave-function curve and encourage more electrons to penetrate. That is, at temperatures above 10 °K, this may be a tunnelling through potential barriers rather than a tunnelling through a slice of ohmic resistor. Or, if you like, frustrated internal reflection.

As to the reason for the shapes of curves (b) and (c), these seem vaguely reconcilable with the resonance picture of superconductivity; for in (b), a mechanism that established resonance between conduction electrons and lattice vibrations would be selective in its reception of electron-waves of other wavelengths, while the hump in (c) could be interpreted as a minor resonance peak. This is just dim speculation, but at least you can see the possibilities that this new discovery has opened for attacking the fundamental physics of the Fermi level; for any explanation must involve the details of the energy distribution of the electrons in the neighbourhood of this.

RADIO INDUSTRY COUNCIL APPOINTMENTS

Mr. J. W. Ridgeway, O.B.E. (A.E.I.), has been elected chairman of the Radio Industry Council in succession to Mr. E. E. Rosen (Ultra).

Mr. A. L. Sutherland (Philips Electrical), who is at present chairman of B.R.E.M.A., has been appointed vice-chairman.

Mr. R. Kelf-Cohen, C.B., who became acting-director of the Radio Industry Council on the death of Sir Raymund Hart last year, has been appointed Director and Secretary, R.I.C.

Information Theory

A DETAILED INTRODUCTION

By P. Holroyd, B.Sc., Grad. Inst. P.* and G. P. Jones, B.Sc., Ph.D.†

Providing, as it does, a fundamental insight into the nature of communication problems, information theory is a valuable tool. Its rapid development over the last few years shows that it has engaged the interest of not only the communications engineer but also of the physicist, the mathematician, the linguist, the biologist, etc., when the complete communication from mind-to-mind is considered.

The theory originated through the work of Wiener, Shannon, Weaver[1,2] and others, and provides means for assessing quantitatively the commodity which is transmitted, namely, the 'information content' of a message, and can be used to show to what extent existing techniques fall short of what is theoretically attainable.

Fig. 1 represents the generalized communication system to be examined from the point of view of information theory. The information source produces the message to be transmitted to the receiver and generally includes a transducer to convert a message (e.g., letters, voice, pictures, temperatures, etc.) into the form of a single-valued time function (for continuous messages) or time series (for discrete messages).

The transmitter encodes or modulates this message into a form suitable for direct transmission over the transmitting medium. In general, it will consist of:

(a) A coder (as an information transformer) which transforms the message with one set of statistical characteristics into a message with a different set of statistical characteristics so that it is matched to the channel allocated to it.

(b) A 'modulator' where the coded message is used to vary the amplitude, frequency, or phase, of a carrier-wave which will then carry the message directly over the transmission medium.

In general, messages from a number of information sources, each being allocated a different channel, will be combined into one signal to modulate the carrier in a multi-channel system.

The waves carrying the coded messages will then arrive at the receiver where they are perturbed by noise, and the inverse operations are carried out to retrieve the original messages as accurately as possible before delivery to the destination.

In dealing with the communication system, the engineer is *not* concerned with the truth or value of the message transmitted but only in conveying it to the destination in the most efficient manner possible. In fact, the information source selects a particular message from a set of messages and the choice must be made known at the receiver. The problem presented to the engineer is one of designing the most efficient system for a given set of conditions and this, as will be shown later, involves proper matching at all points in the system of the quantity called information, just as impedances should be matched in electrical networks.

We shall now develop the important theorems of information theory and proceed with simple examples to show how they are used in practice.

Definition of Information

In general, any message to be transmitted will be made up of symbols selected at discrete moments from a collection or ensemble of symbols. Knowing the total number of symbols in the message (n) and the number of different symbols available for selection (s) it is seen

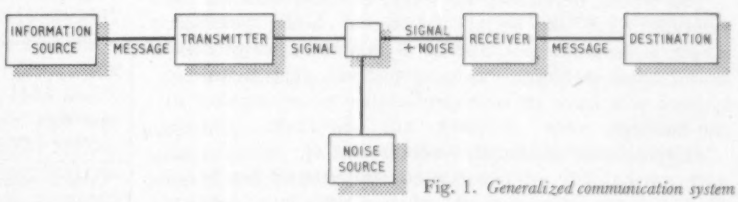


Fig. 1. *Generalized communication system*

that the possible number (m) of different messages of length (n) will be given by

$$m = s^n$$

If the information source selects a message purely at random from the total number available, each message will have an equal probability of being selected, given by $1/m$.

Since this represents the degree of choice available to the sender in selecting a message for transmission it must in some way represent the *selective information content* of that message. We define, therefore, the selective information content of a message as some function (F) of its probability of being selected from all possible messages. Intuitively in this case, since all the possible messages are equally probable, each must contain the same amount

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of information, and obviously two messages should contain twice as much information, three messages three times as much, and so on. Our function (F) must therefore possess this additive property.

Probability of selecting x messages is $(1/m)^x$. Information content of x messages is $F(1/m)^x \equiv xF(1/m)$. Let $1/m$ equal a, then $F(a)^x \equiv xF(a)$.
Differentiating with respect to a obtaining

$$F'(a^x)\ x\ a^{x-1} = xF'(a)$$
$$\text{or} \quad a^x F'(a^x) = aF'(a)$$
$$\text{therefore} \quad aF'(a) = \text{constant } (k)$$
$$\text{hence} \quad F'(a) = k/a$$
$$\text{or} \quad F(a) = k \log(a) + K$$

Thus, a function which satisfies the identity $F(a^x) \equiv x\,.\,F(a)$ is the logarithm and, in order to make the information content positive, we choose ($-$log). Therefore, the information content of a message is $-\log(1/m)$ or $\log(m) = n \log(s)$. This formula was first introduced by R. V. L. Hartley in 1927 in a paper to the International Congress of Telegraphy and Telephony.[3]

The simplest message would consist of one symbol ($n = 1$) selected from two equally probable symbols ($s = 2$) and the information content would be log2. This is taken to represent one unit of information and, hence, the logarithm must be taken to base two, the unit is then called a 'binary digit', usually referred to as a 'bit'.

For example, if the message 'AAABBC' is selected from the set of all possible combinations of A, B, C, taken six at a time the information content is:

$$\log_2(3^6) = 6 \log_2(3) = 9{\cdot}51 \text{ bits}$$

The information content of the message—'Weak, without milk or sugar please', taken from the set-weak-strong, with and without milk and sugar, is given by $\log_2(6^3)$ or $3 \log_2(6) = 7{\cdot}75$ bits.

The above definition, however, assumes that all the symbols have an equal chance of being selected, which is certainly not true in a great number of communication problems. In general then, each available symbol will have its own probability of occurrence in the message.

Suppose our available symbols are $x_1, x_2, \ldots x_m$, with probability of occurrence represented by $p(x_1)$, $p(x_2) \ldots p(x_m)$ respectively, then a very long message of N symbols will contain $p(x_1)$ N symbols of type x_1, $p(x_2)$ N symbols of type x_2 and so on, so that the probability of occurrence of the extremely long message is

$$p(N) = [p(x_1)]^{p(x_1)N} \times [p(x_2)]^{p(x_2)N} \times \ldots [p(x_m)]^{p(x_m)N}$$

and its information content according to our definition is

$$I(N) = -\log_2 p(N) = -N \sum_{x=x_1}^{x=x_m} p(x) \log_2 p(x)$$

So that the average information per symbol is

$$I(x) = -\sum_x p(x) \log_2 p(x). \quad .. \quad .. \quad .. \quad (1)$$

This general definition for the selective information content of a message based on the statistical selection of symbols was derived by C. E. Shannon in 1948[4].

If we have s symbols all equally probable then $p(x) = 1/s$ and $I(x) = \log_2(s)$ so that $I(n) = n \log_2(s)$ in agreement with the Hartley formula. In fact, when the symbols are equally probable, the information content per symbol is a maximum.

This statistical measure of information is important because it determines the saving in transmission time that is possible by proper encoding of a message making use of its statistical characteristics.

For instance, we may wish to determine the efficiency of a code that is used to transmit information obtained from a language of given letter probabilities. Consider, for example, a five-letter language with the letter probabilities given in Table 1.

TABLE 1

Letter	A	B	C	D	E
Letter Probability ..	1/2	1/16	1/4	1/16	1/8

Average information per letter
$-\frac{1}{2}\log_2\frac{1}{2} - 1/16\log_2(1/16) - \frac{1}{4}\log_2\frac{1}{4} - 1/16\log_2(1/16)$
$-\frac{1}{8}\log_2\frac{1}{8} = 1{\cdot}88$ bits per letter.

Suppose we require that the messages be transmitted as a series of on/off pulses over an electrical communication system; i.e., as a series of binary digits (0 or 1). It is obvious that the code used should allocate the least number of digits to the most probable letter, with increasing complexity for letters of lower probability. To show this, a comparison of two codes will be made, one of the codes being devised independently of the letter probabilities, while the other will be approximately matched to the letter probabilities.

The two codes are given in Table 2.

TABLE 2

Letter	A	B	C	D	E
Code 1 (unmatched) ..	0	10	110	1110	1111
Code 2 (matched) ..	0	1110	10	1111	110
Letter Probability ..	1/2	1/16	1/4	1/16	1/8

The message ADDCEBDCA would become

01110111011011111011101100 in code 1 and

01111111111011011101111100 in code 2.

Consider code 1. In a long message of N letters we shall have, when coded, the following number of digits

No. of 0s $= N(1/2 + 1/16 + 1/4 + 1/16) = (14/16)N$
No. of 1s $= N(1/16 + 2/4 + 3/16 + 4/8) = (20/16)N$

Total number of digits $= (34/16)N$
So that the probability of a 0 is 7/17, and the probability of a 1 is 10/17.

Thus, the average information per digit is

$-7/17 \log_2(7/17) - 10/17 \log_2(10/17)$

$= 0{\cdot}98$ bit/digit.

Average information per letter $0{\cdot}98 \times 34/16 = 2{\cdot}07$ bits/letter.

Consider code 2. In this case, a long message of N letters will give

No. of 0s $= N(1/2 + 1/16 + 1/4 + 1/8) = (15/16)N$

No. of 1s $= N(3/16 + 1/4 + 4/16 + 2/8) = (15/16)N$

Total number of digits $= (30/16)N$ and the 0s and 1s are equally probable, so that the average information per digit $-\frac{1}{2}\log_2(\frac{1}{2}) - \frac{1}{2}\log_2(\frac{1}{2}) = 1{\cdot}0$ bit/digit. Average information per letter $1{\cdot}0 \times 30/16 = 1{\cdot}88$ bits/letter.

Hence, for this particular code, the letter probabilities match perfectly, and this code is obviously the most efficient to use for this case.

So far, we have considered only the encoding of discontinuous messages in the form of discrete letters but, in certain circumstances, it is necessary to be able to encode a continuous message such as the waveforms produced by speech, or any quantity whose value is known to vary with time.

To do this, we make use of a theorem due to J. Carson[5] known as the 'sampling theorem'. If a function $G(t)$ contains no higher frequencies then W c/s, then it is completely determined by giving its ordinates at a series of points spaced $W/2$ seconds apart, the series extending throughout the time domain.

This means that $2WT$ samples of a waveform are sufficient to specify it completely. Thus to encode a continuous waveform of highest frequency W c/s samples are taken at $2WT$ points, and the sample magnitudes encoded, this is then sufficient for the receiver to reconstruct the message waveform completely.

Since the sample magnitudes may have one of an infinite number of values, any finite encoding must limit the accuracy of the sample. For example, a sample magnitude may in fact be 6.746652103869337 volts which, if we could measure to such accuracy, would require approximately 68 bits of information for its transmission (each digit requires 4 bits, $1111 \equiv 15$, if we use a binary code). If less than this are used then there will be some inaccuracy in the reconstructed magnitude at the receiver. In practice, no attempt is made to achieve such extravagant accuracy and it is usual to divide the possible variation of the waveform into Q sections, where Q is a number of the form 2^n, then only the number of divisions nearest to the sample magnitude is transmitted. This procedure is known as amplitude quantizing the waveform.

The waveform in Fig. 2 has been sampled at points $W/2$ apart and the value of the function at these points quantized into eight levels. For higher accuracy, a larger number of quantizing levels must be used. In this example the message to be transmitted is

2, 3, 3, 2, 1, 2, 1, −2, −3, −3, −3, −2, −1, −1, −1, −1,

and this may be encoded in the same way that the simple language was encoded above. If each decimal digit is converted to the equivalent binary expression

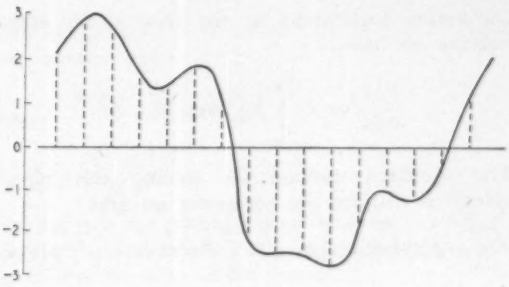

Fig. 2. *Amplitude quantizing; the waveform has been sampled at points $W/2$ apart and the value of the function at these points quantized into eight levels*

then, as each decimal requires three binary digits (there being only eight levels; i.e., 2^3), the complete waveform would require $3 \times 16 = 48$ bits for its transmission.

It will be obvious that since we have quantized the sample magnitudes of the correct waveform at the source, it is not possible to reconstruct the waveform correctly at the receiver.

The reconstructed waveform will be distorted from the original and this is equivalent to noise affecting the signal during transmission. The effect is, therefore, referred to as quantizing noise, and it decreases with increasing number of quantizing levels[6].

In the case of a continuous message, it is apparent that the information content per sample can be obtained by replacing the discrete probabilities $p(x_1)$ $p(x_2)$ $p(x_3)$, etc. by a continuous probability density function; and by replacing the summation by integration over the whole range of possible amplitudes; i.e.,

$$I(x) = -\int_x p(x) \log_2 p(x)\, dx \qquad .. \qquad .. \qquad .. \quad (2)$$

If the message waveform is of duration T seconds, and bandwidth W c/s then the number of samples required is $2WT$; so that the information content of the complete waveform is—

$$I(2WT) = 2WT\left[-\int_x p(x) \log_2 p(x)\, dx\right]$$

assuming that all the samples are statistically independent. It follows that the information content transmitted or received per second is

$$I = -2W \int_x p(x) \log_2 p(x)\, dx$$

It is necessary when dealing with practical systems to impose some restriction upon the unknown distribution $p(x)$ and to determine what form of $p(x)$ makes the information content a maximum.

Let us consider a case in which the mean squared value of the amplitude at the sample points is fixed; i.e., the mean power P (say) is fixed. We have to maximize the integral

$$I(x) = -\int_{-\infty}^{+\infty} p(x) \log_2 p(x)\, dx$$

subject to the conditions

$$P = \int_{-\infty}^{+\infty} x^2 p(x)\, dx \text{ and } 1 = \int_{-\infty}^{+\infty} p(x)\, dx$$

Converting logarithms to the base e, for ease of calculation we have:

$$\frac{I(x)}{\log_2 e} = -\int_{-\infty}^{+\infty} p(x)\,\log_e p(x)\,dx$$

The standard method for solving this type of problem[7] is to form the composite integral

$$\int_{-\infty}^{+\infty} -p(x)\log_e p(x)\,dx + \lambda\int_{-\infty}^{+\infty} x^2 p(x)\,dx + \mu\int_{-\infty}^{+\infty} p(x)\,dx$$

or $$\int_{-\infty}^{+\infty} p(x)\,[-\log_e p(x) + \lambda x^2 + \mu]\,dx$$

where λ and μ are undetermined multipliers, and to maximize by varying $p(x)$; i.e., when $\frac{\partial}{\partial p} = 0$. Thus we have,

$$-1 - \log_e p(x) + \lambda x^2 + \mu = 0$$

or $$p(x) = e^{\mu-1}\exp(\lambda x^2)$$

Using the given conditions we have

$$\int_{-\infty}^{+\infty} x^2 e^{\mu-1}\exp(\lambda x^2)\,dx = P$$

$$\text{or } e^{\mu-1}\frac{1}{2(-\lambda)}\sqrt{\frac{\pi}{(-\lambda)}} = P$$

$$\text{and } \int_{-\infty}^{+\infty} e^{\mu-1}\exp(\lambda x^2)\,dx = 1$$

$$\text{or } e^{\mu-1}\sqrt{\frac{\pi}{(-\lambda)}} = 1$$

Therefore, $\lambda = -\frac{1}{2P}$ and $e^{\mu-1} = \frac{1}{\sqrt{2\pi P}}$

Hence $p(x) = \frac{1}{\sqrt{2\pi P}}\exp\left(-\frac{x^2}{2P}\right)$

i.e., a Gaussian distribution of amplitude. The information content will be—

$$\frac{I(x)_{(max)}}{\log_2 e} =$$

$$-\frac{1}{\sqrt{2\pi P}}\int_{-\infty}^{+\infty}\exp\left(-\frac{x^2}{2P}\right)\log_e\left[\frac{1}{\sqrt{2\pi P}}\exp\left(-\frac{x^2}{2P}\right)\right]dx$$

$$= -\frac{1}{\sqrt{2\pi P}}\int_{-\infty}^{+\infty}\exp\left(-\frac{x^2}{2P}\right)\left[\log_e\left(\frac{1}{\sqrt{2\pi P}}\right) - \frac{x^2}{2P}\right]dx$$

but $$\int_{-\infty}^{+\infty}\exp\left(-\frac{x^2}{2P}\right)dx = \sqrt{2\pi P}\text{ and}$$

$$\int_{-\infty}^{+\infty}\frac{x^2}{2P}\exp\left(-\frac{x^2}{2P}\right)dx = \tfrac{1}{2}\sqrt{2\pi P}$$

Therefore $$\frac{I(x)_{max}}{\log_2 e} = -\left[\log_e\frac{1}{2\pi P} - \tfrac{1}{2}\right]$$

$$= -\left[\log_e\frac{1}{\sqrt{2\pi P}} + \log_e\frac{1}{\sqrt{e}}\right]$$

or $$\frac{I(x)_{max}}{\log_2 e} = -\log_e\frac{1}{\sqrt{2\pi eP}} = \log_e\sqrt{2\pi eP} = \tfrac{1}{2}\log_e 2\pi eP.$$

$$\therefore I(x)_{max} = \tfrac{1}{2}\log_2 e\,\log_e(2\pi eP) = \tfrac{1}{2}\log_2(2\pi eP) \quad . . \quad (3)$$

Examples of other types of constraint on the waveform and the resulting probability distribution are given in Table 3.

TABLE 3

Type	Limitation	Prob. Distr.	Inf. Cont.
Symmetrical Signal Distribution	mean power	$p(x) = \frac{1}{\sqrt{2\pi P}}\,e^{-\frac{x^2}{2P}}$	$\frac{1}{2}\log_2 2\pi eP$
	peak power limitation $\hat{P}$ at sampling points	$p(x) = \frac{1}{2\sqrt{\hat{P}}}$	$\frac{1}{2}\log_2 4\hat{P}$
	peak power limitation $\hat{P}$ throughout time		$\frac{1}{2}\log_2\frac{4\hat{P}}{e^2} > Ix > \frac{1}{2}\log_2 4\hat{P}$
Unidirectional Signal Distribution	x positive average A_0	$p(x) = \frac{1}{A_0}\exp\left(-\frac{x}{A_0}\right)$	$\log_2 e\,A_0$
	average P power	$p(x) = \sqrt{\frac{2}{\pi P}}\,e^{-\frac{x^2}{2P}}$	$\frac{1}{2}\log_2(\frac{1}{2}\pi e P)$

Reception

During transmission, the signal is subject to 'noise' from various sources; the nature of which is irrelevant to the present discussion. The main effect of the noise is to change the transmitted signal. The observer then has the problem of deciding the probability that the received message y say represents the particular transmitted message x.

For example, suppose we receive the message

'I WILL MCET TOU NONIGHT ASEARRANGLD'.

This is the received message y and there is little difficulty in deciding that the probability of the transmitted message x being

'I WILL MEET YOU TONIGHT AS ARRANGED'

is high.

The fact that we can make this type of correction to written English means, of course that, as a code, English is very inefficient; more information is transmitted than is required by the observer to decode the message. This is obvious from the fact that the error did not affect the recognition of the true message. Such an effect is called 'redundancy' and an efficient code should have as little redundancy as possible. The perfect code should have no redundancy and, under these conditions, any error in the received signal could not possibly be recognized.

Redundancy is, however, sometimes specially used to overcome the effects of noise on the signal. For instance, if the noise is very large and on transmitting a

sequence of Q letters a certain percentage of Q are in error at the receiver, then it is possible to combat the noise by repeating the information a sufficient number of times; this, in fact, is our normal reaction to a message that is not fully understood when listening to a person speaking.

As an example of a message badly distorted due to noise, consider Table 4.

TABLE 4

Message 'y'	DPCHRXPOSAGFTISREPEMIER
Repeat 1	FGCHHMETSAGEOINWRDPECTED
Repeat 2	EANHMEIOASEIFREPCSKDP
Repeat 3	EAINMESTOYEIDRHPUATAR
Repeat 4	NYCHFSEDSAGEESRRMPTATEC
Repeat 5	WLEHBLSSAREIORPMEATED
–and so on	

As each message is repeated the probability of each letter being correct is either increased or decreased, thus with sufficient number of repetitions the original message may be retrieved. In the example given in Table 4, by writing down the letter which occurs most frequently in each column the original message is, to a high degree of probability, recovered. This is a particular example of a general case which will now be considered in some detail.

In a practical communications system, the presence of noise will make it impossible for the observer at the receiver to identify with certainty which symbol was transmitted from the source. In other words, the 'aposteriori probability' for a transmitted symbol x will not, in general, be unity but $p_y(x)$ say which, in fact, is the conditional probability that the transmitted symbol was x on receiving the signal y.

The way in which this affects the measure of information gain $(-\log p(x))$ can be seen if we allow the symbol x to be transmitted twice; first in the presence of noise, and secondly under ideal noise-free conditions. During the first transmission, the probability at the receiver changes from $p(x)$ to $p_y(x)$ and, therefore, after the second trial it must have changed from $p_y(x)$ to unity. The total gain of information is $[-\log p(x)]$, while the gain from the second transmission is $-\log p_y(x)$. Hence, using the additive property the gain from the first transmission must have been

$$-\log p(x) - [-\log p_y(x)]$$

or the information gain is

$$\log \frac{p_y(x)}{p(x)} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (4)$$

which gives a general definition of information gain in the presence of noise.

By the product law of probabilities, the joint probability $p(x,y)$ that a value of x and a value of y will occur together is given by

$$p(x,y) = p(x)\, p_x(y) = p(y)\, p_y(x)$$

so that

$$p_y(x) = \frac{p(x)\, p_x(y)}{p(y)} \quad .. \quad .. \quad .. \quad .. \quad (5)$$

where $p_x(y)$ is the probability of receiving a y when x is transmitted and consequently specifies the statistical properties of the noise in the channel.

If we receive a certain y, Equ. (5) can be used to assess the relative aposteriori probability $p_y(x)$ of which symbol x was the transmitted one. The observer will never be absolutely certain which symbol was transmitted and so will be unable to express the gain of information by means of Equ. (4) as such, However. it would seem reasonable to state that, whatever the transmitted symbol, the same received y should represent the same quantity of received information. Thus Equ. (4) is averaged over all values of x to give the average information gain when a y is received.

In a long received message of N samples, y_1 will be received $p(y_1) \cdot N$ times; of these $p_{y1}(x_1)\, p(y_1)\, N$ will be due to x_1 having been transmitted. $p_{y1}(x_2)\, p(y1)N$ will be due to x_2 having been transmitted and so on.

The total gain in information due to all the received y_1s will then be

$$p(y_1)\, N P_{y1}(x_1) \left[\log_2 \frac{p_{y1}(x_1)}{p(x_1)}\right] + p(y_1)\, N p_{y1}(x_2) \left[\log_2 \frac{p_{y1}(x_2)}{p(x_2)}\right] + \ldots$$

$$= p(y_1)\, N \sum_x p_{y1}(x) \log_2 \frac{p_{y1}(x)}{p(x)}$$

Average information gain when y_1 is received is given by

$$\text{Av.}_x\, I(y_1) = \sum_x p_{y1}(x) \log_2 \frac{p_{y1}(x)}{p(x)}$$

Similarly, total gain of information due to all the received y_2s will be

$$p(y_2)\, N \sum_x p_{y2}(x) \log_2 \frac{p_{y2}(x)}{p(x)}$$

Hence the total information due to the reception of N samples will be

$$I(N) = p(y_1)\, N \sum_x p_{y1}(x) \log_2 \frac{p_{y1}(x)}{p(x)} + p(y_2)\, N \sum_x p_{y2}(x) \log_2 \frac{p_{y2}(x)}{p(x)} + \ldots$$

$$= N \sum_y p(y) \sum_x p_y(x) \log_2 \frac{p_y(x)}{p(x)}$$

and the average information per sample is

$$I = \sum_y p(y) \sum_x p_y(x) \log_2 \frac{p_y(x)}{p(x)}$$

This can be written

$$I = \Sigma_y\, p(y) \{-\Sigma_x\, p_y(x) \log_2 p(x)\} - \Sigma_y\, p(y) \cdot \{-\Sigma_x\, p_y(x) \log_2 p_y(x)\}$$

and, by means of Equ. (5), the first term on the right-hand side will become

$$\sum_y p(y) \left\{ -\sum \frac{p_x(y)}{p(y)}\, p(x) \log_2 p(x) \right.$$

$$= p(y_1) \left\{ -\frac{p_{x1}(y_1)}{p(y_1)}\, p(x_1) \log_2 p(x_1) \right.$$

$$\left. - \frac{p_{x2}(y_1)}{p(y_1)}\, p(x_2) \log_2 p(x_2) - \ldots \right\}$$

$$+ p(y_2) \left\{ -\frac{p_{x1}(y_2)}{p(y_2)}\, p(x_1) \log_2 p(x_1) \right\}$$

$$\left. - \frac{p_{x2}(y_2)}{p(y_2)}\, p(x_2) \log_2 p(x_2) - \ldots \right\}$$

$$+ \text{etc.}$$

$$= -p(x_1) \log_2 p(x_1) \sum_y p_{x1}(y)$$

$$- p(x_2) \log_2 p(x_2) \sum_y p_{x2}(y) - \ldots$$

$$= -p(x_1) \log_2 p(x_1) - p(x_2) \log_2 p(x_2) - \ldots$$

(Since $\sum_y p_x(y) = 1$)

$$= -\sum p(x) \log_2 p(x)$$

$= I(x)$ [from Equ. (1)]

The second term may be written

$$p(y_1) \{ -\textstyle\sum_x p_{y1}(x) \log_2 p_{y1}(x) \}$$

$$+ p(y_2) \{ -\textstyle\sum_x p_{y2}(x) \log_2 p_{y2}(x) \} + \ldots$$

which is the average over all values of y with $p(y)$ as the weighting function of the term

$$-\textstyle\sum_x p_y(x) \log_2 p_y(x) = I_y(x) \text{ say.}$$

Hence, the information gain per sample is

$$I_x - \text{Av.}_y\, I_y(x) \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

Since there is complete symmetry between x and y (viz., Equ. 5) we may also write

$$I = I_y - \text{Av.}_x\, I_x(y) \quad .. \quad .. \quad .. \quad .. \quad (7)$$

The continuous case may, as before, be considered by replacing the discrete probability density functions and the summations by integrals.

In order to illustrate the use of these equations, let us consider one or two examples in which a number of messages are transmitted, affected by noise, and received as a new set of messages.

Suppose a sequence of eight simple messages is transmitted, these being

YES, YES, YES, YES, YES, NO, NO, NO.

TABLE 5

x	YES	YES	NO	NO
y	UP	DOWN	UP	DOWN
$p(x,y)$	1/4	3/8	1/8	1/4
$p(x)$	5/8	5/8	3/8	3/8
$p(y)$	3/8	5/8	3/8	5/8
$p_y(x)$	2/3	3/5	1/3	2/5
$p_x(y)$	2/5	3/5	1/3	2/3
$\text{Log}_2 p_y(x)/p(x)$	+0·09	−0·06	−0·17	+0·09

The receiver can be a device (i.e., a lever) that indicates a YES by being up and a NO by being down but, due to noise, the following sequence is received:
UP, UP, DOWN, DOWN, DOWN, DOWN, DOWN, UP.
We can now set out a table of probabilities (Table 5).

The figures in the last row give the information gain for the four combinations of x and y. A negative information gain means that we have been misled; we are now worse off in our knowledge of the true state of affairs than we were before receiving the message.

The average information received is of more interest and it has been shown above that this is given by

$$\text{Av}_x\, I(y) = \textstyle\sum_x p_y(x) \log_2 p_y(x)/p(x)$$

$$\text{Av}_y\, I(x) = \textstyle\sum_y p_x(y) \log_2 p_y(x)/p(y)$$

$$I = \textstyle\sum_y p(y) \sum_x p_y(x) \log_2 p_y(x)/p(x)$$

$$= \sum_y \sum_x p(x,y) \log_2 \frac{p(x,y)}{p(x) \cdot p(y)}\,.$$

Thus, in the above examples, we have

I (up) $= (2/3)\ (0{\cdot}09) + (1/3)\ (-0{\cdot}17) = 0.0033$ bit

I (down) $= (3/5)\ (-0{\cdot}06) + 2/3\ (0{\cdot}09) = 0{\cdot}0240$ bit

I (YES) $= (2/5)\ (0{\cdot}09) + 3/5\ (-0{\cdot}06) = 0$ bit

I (NO) $= (1/3)\ (-0{\cdot}17) + 2/5\ (0{\cdot}09) = -0{\cdot}027$ bit

$I = (1/4)\ (0{\cdot}09) + 1/8\ (-0{\cdot}17) + 3/8\ (-0{\cdot}06) +$
$1/4 \times 0{\cdot}09 = 0{\cdot}00125$ bit/symbol.

The fact that these quantities are very small is due to the large amount of 'noise' present between the transmitter and the receiver. A much less noisy case is given below.

Let the message sequence transmitted be:

x 0, 0, 0, 0, 0, 0, 1, 1, 1, 1,

and the received sequence be:

y a, a, a, a, a, b, b, b, b, b,

Then we have the probabilities shown in Table 6.

TABLE 6

x	0	0	1	1
y	a	b	a	b
$p(x,y)$	1/2	1/10	0	2/5
$p(x)$	3/5	3/5	2/5	2/5
$p(y)$	1/2	1/2	1/2	1/2
$p(x)$	1	1/5	0	4/5
$p_x(y)$	5/6	1/6	0	1
$\log_2 \frac{p_y(x)}{p(x)}$	0·74	−1·58	$-\infty$	1·0

Then, as before

$I(0) = (5/6)\ (0{\cdot}74) - (1/6)\ (1{\cdot}58) = 0{\cdot}342$ bit

$I(1) = (0)\ (-\infty) + (1)\ (1) = 1{\cdot}000$ bit

$I(a) = (1)\ (0{\cdot}74) + (0)\ (-\infty) = 0{\cdot}740$ bit

$I(b) = (1/5)\ (-1{\cdot}58) + 4/5\ (1{\cdot}0) = 0{\cdot}484$ bit

$I = (1/2)\ (0{\cdot}74) - 1/10\ (1{\cdot}58) + 0 + 2/5\ (1{\cdot}0)$
$= 0{\cdot}612$ bit/digit

If there had been no errors (i.e., no noise) the number of bits per digit transmitted would have been

$$I(x) = -\sum p(x) \log_2 p(x)$$
$$= -6/10 \log_2 6/10 - 4/10 \log_2 4/10$$
$$= 0{\cdot}972 \text{ bit/digit}$$

whereas, in actual fact, only 0·612 bit/digit were transmitted.

These examples are, of course, ideal in that a knowledge of the transmitted message is not usually available at the receiver; indeed, the actual transmission would be pointless if we already knew the message. However, this procedure when applied to actual systems, by postulating transmitted-message noise, and received-message probability functions, enables one to deduce the theoretical rate of transmission of information and define the 'channel capacity' in each case.

The rate of transfer of information on the channel can be defined as the average gain of information per unit time at the destination. Suppose the channel has a bandwidth W c/s and that it is used for a time T seconds, then any functions of time whose spectra lie entirely within the band W and time functions within T can be used to convey information over the channel. The message to be transmitted is thus converted into these signal functions at the transmitter. Each signal function can be completely described by $2WT$ amplitude samples spaced $W/2$ seconds apart throughout the time period T and, if there is no noise whatever on the channel, it is possible to distinguish an infinite number of different amplitude levels at each sampling point. We could, therefore, distinguish an infinite number of different signal functions on the channel and, consequently, the channel capacity would be infinite. Even if there is noise present, an infinite number of different signal functions could be distinguished if the transmitter had an unlimited power. Only when the signal power is limited in some way and noise is present does there exist a finite capacity for the channel.

Let $x(t)$ be the transmitted signal, $n(t)$ the noise in the channel and $y(t)$ the received signal. Suppose the noise is simply added to the signal so that

$$y(t) = x(t) + n(t)$$

We now wish to decide what signals $x(t)$ should be transmitted to make the rate of transfer of information along the channel a maximum; and what is the value of this maximum rate when certain conditions are imposed on the signal.

As an example, suppose we have white gaussian noise on the channel; that is, noise with a uniform frequency spectrum and gaussian amplitude distribution and suppose, also, that the signal is limited in mean power, then $\overline{x^2} = P$ (mean signal power) $\overline{n^2} = N$ (mean noise power).

The mean transfer of information per sample is given by Equ. (7)

$$I = I_y - \text{Av.}_x I_x(y)$$

Now $I_x(y)$ depends only on $p_x(y)$ which, in time, is the probability distribution of the noise n. Therefore, we may write

$$I = I_y - I_n$$

But, from Equ. (3) $I_n = \frac{1}{2} \log_2 2\pi e N$, since the mean squared value of the amplitude is fixed and it has a gaussian distribution. It was also shown that Equ. (3) represents the maximum value possible with the restriction that the mean squared power is fixed. Hence in order to maximize the rate of transmission of information the received signal $y(t)$ must be of the same nature as white gaussian noise. It follows also that the transmitted signals $x(t)$ must be of the same nature as white gaussian noise so that the messages to be transmitted must be coded into signals that resemble white gaussian noise as closely as possible.

By analogy with Equ. (3)

$$I_y = \tfrac{1}{2} \log_2 2\pi e (P + N)$$

Therefore

$$I_{max} = \tfrac{1}{2} \log_2 2\pi e (P + N) - \tfrac{1}{2} \log_2 2\pi e N$$
$$= \tfrac{1}{2} \log_2 \frac{P + N}{N}$$

In time T seconds there are $2WT$ samples so that the total gain of information in time T is

$$I(T)_{max} = WT \log_2 \frac{P + N}{N}$$

Therefore, maximum information rate R_{max} is

$$R_{max} = W \log_2 \left(\frac{P + N}{N} \right) \quad .. \quad .. \quad .. \quad (9)$$

This is Shannon's[4] mean power theorem—one of the most important results of information theory. By sufficiently complicated encoding it is possible to transmit information over the channel at this maximum rate with as small a frequency of errors as desired. This rate cannot be exceeded by any encoding method and still have an arbitrarily low frequency of errors, so that R_{max} may be described as the capacity of the channel in bits/sec (when the logarithm is of base two) for the transmission of information.

Another example is provided by having an arbitrary type of noise in the channel, the signal still having a mean power limitation P. Let N be the mean power of the noise, and $\overline{N}$ be the mean power of the white gaussian noise signal limited to the same bandwidth and time, and having the same information content as the arbitrary noise. Since, as has been proved, white gaussian noise has the maximum $I(x)$ of any ensemble for a given mean power, for any ensemble, signal or noise $\overline{N} \leqslant N$.

The mean power of the received signal is $(P + N)$ and, hence, its information content cannot exceed $WT \log_2 2\pi e (P + N)$, since this would be the value if its distribution were that of white gaussian noise. By definition, the information content of the noise is

$$WT \log_2 2\pi e \overline{N}$$

so that

$$I(T) \leqslant WT \log_2 2\pi e (P + N) - WT \log_2 2\pi e \overline{N}$$
$$= WT \log_2 \left(\frac{P + N}{\overline{N}} \right)$$

If the transmitted signal is equivalent to white gaussian noise, then the information content of the

signal will be at least as great as $WT \log_2 2\pi e (P + \overline{N})$

so that $$I(T) \geqslant WT \log_2 \frac{P + \overline{N}}{\overline{N}}$$

Thus we have an upper and lower limit to the maximum rate of transfer of information in a noisy channel

$$W \log_2 \left(\frac{P+N}{\overline{N}}\right) \geqslant R_{max} \geqslant W \log_2 \left(\frac{P+\overline{N}}{\overline{N}}\right) \quad (10)$$

If $P/\overline{N}$ is large then, $R_{max} = W \log_2 \frac{P}{\overline{N}}$.

From these examples we see that the general equation for the gain of information at the receiver (Equs. 6 and 7) can be used to evaluate, or at least set upper and lower limits to the rate of transmission of information from one point to another.

In this way various transmission systems may be compared with each other or with an ideal transmission system from the point of view of their usefulness and efficiency for transmitting information.

Noise Reduction

We have already mentioned that one method of combating noise is to make use of redundancy. The message, or part of the message, is repeated a sufficient number of times to bring the probability of the transmitted message being the received message to a sufficiently high value.

Other methods of noise reduction using redundancy are available; an excellent example being the error detecting and correcting codes devised by Hamming[8]. In this type of code a certain number of digits are used, not to carry information, but to provide a means of checking the code at the receiver. For example, if we have seven digits then three of these must be used for checking purposes, since $2^3 = 8$; i.e., 0 to 7 inclusive.

Let the four information digits be $x_1\ x_2\ x_3\ x_4$ and the three correcting digits be $x_5\ x_6\ x_7$. A number $\alpha\ \beta\ \gamma$ will be formed such that $\alpha\ \beta\ \gamma$ gives the position of a single error if any.

For an error in the 4th, 5th, 6th or 7th position we must have $\alpha = 1$

$$\therefore \quad \alpha = x_4 + x_5 + x_6 + x_7$$

For no errors $\alpha = 0$, therefore $x_4 + x_5 + x_6 + x_7 = 0$

or $\quad x_7 = x_4 + x_5 + x_6$

For an error in the 2nd, 3rd, 6th or 7th position we must have $\beta = 1$

$$\therefore \quad \beta = x_2 + x_3 + x_6 + x_7$$

For no errors $\beta = 0$, therefore $x_2 + x_3 + x_6 + x_7 = 0$

or $\quad x_6 = x_2 + x_3 + x_7$

For an error in the 1st, 3rd, 5th or 7th position, we must have $\gamma = 1$

$$\therefore \quad \gamma = x_1 + x_3 + x_5 + x_7$$

For no errors $\gamma = 0$, therefore $x_1 + x_3 + x_5 + x_7 = 0$

or $\quad x_5 = x_1 + x_3 + x_7$

Thus $\quad x_6 = (x_2 + x_3) + (x_4) + x_5 + x_6$

$$\left.\begin{array}{lll} \therefore & x_5 = & x_2 + x_3 + x_4 \\ \text{similarly} & x_6 = & x_1 + x_3 + x_4 \\ \text{hence} & x_7 = & x_1 + x_2 + x_4 \end{array}\right\} \quad .. \quad (A)$$

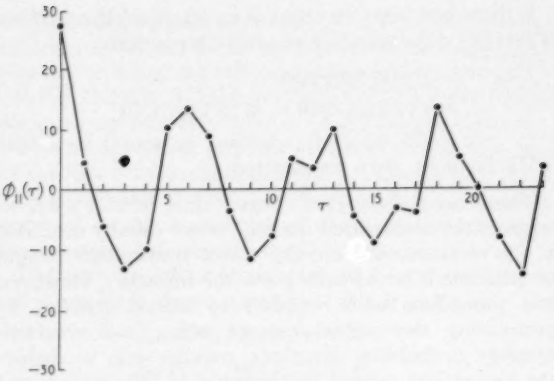

Fig. 3. *Auto-correlogram*

Example. Suppose the message we wish to transmit is 1011 then, from the above equations, we calculate

$$x_5 = 0 + 1 + 1 = 0$$
$$x_6 = 1 + 1 + 1 = 1$$
$$x_7 = 1 + 0 + 1 = 0$$

hence we transmit the message 1011010. The receiver will be able to detect and correct one error in the received message.

Suppose the received message is 1011110 then, using the above equations for $\alpha\ \beta$ and γ, we find

$$\alpha = 1 + 1 + 1 + 0 = 1$$
$$\beta = 0 + 1 + 1 + 0 = 0$$
$$\gamma = 1 + 1 + 1 + 0 = 1$$

therefore, $\alpha\ \beta\ \gamma$ is equal to 101 which means that the fifth digit is in error. The transmitted signal is thus 1011010 which is true.

For greater correcting power more redundancy is required; i.e., a message of fifteen bits will have four bits redundant.

As well as the use of redundancy in combating noise, other methods of signal detection in the presence of noise are available and one of the most powerful and important of these is the correlation method.[9] It will be convenient to state the correlation function first. and describe its meaning and use later.

If we have two functions of time $y_1(t)$ and $y_2(t)$, then we define the cross correlation function of $y_1(t)$ and $y_2(t)$ as $\phi_{12}(\tau)$ where

$$\phi_{12}(\tau) = \operatorname*{Lim}_{T \to \infty} \frac{1}{2T} \int_{-T}^{+T} y_1(t)\, y_2(t+\tau)\, dt \quad .. \quad (11)$$

where τ is a time delay imposed on one of the functions.

A graph of $\phi_{12}(\tau)$ against τ is called a correlogram or correlation diagram, and it is the correlogram that enables one to detect the presence of signals in noise. If $y_1(t) = y_2(t)$ then the correlation function is known as the auto-correlation function and is

$$\phi_{11}(\tau) = \operatorname*{Lim}_{T \to \infty} \frac{1}{2T} \int_{-T}^{+T} y_1(t)\, y_1(t+\tau)\, dt \quad .. \quad (12)$$

Thus, to obtain a point on the auto-correlation diagram for $\tau = \tau_1$, say, the function $y_1(t)$ is multiplied for all t by the value of $y_1(t)$ delayed by a time of τ_1.

The integral of the product is then taken over the complete duration of the function and the result divided by the duration to give $\phi_{11}(\tau_1)$. This procedure is carried out for all values of τ and the resulting curve is the auto-correlogram.

If a periodic signal is completely immersed in noise then, by following the above procedure, the output from the correlator will consist of a periodic waveform of the same frequency as the hidden signal. To show this consider a signal $y(t)$ given by

$$y(t) = A \cos \omega t + n(t)$$

where $n(t)$ is random noise linearly added to the transmitted signal $A \cos \omega t$.

The auto-correlation function for this case will be

$$\phi_y(t) = \operatorname*{Lim}_{T \to \infty} \frac{1}{2T} \int_{-T}^{+T} [A \cos \omega t + n(t)] \, [A \cos \omega(t + \tau) + n(t + \tau)] \, dt$$

Since the noise and signal are incoherent, terms such as $n(t) \times A \cos \omega t$ vanish, and the above equation reduces to

$$\phi_y(t) = \frac{A^2}{2} \cos \omega\tau + \operatorname*{Lim}_{T \to \infty} \frac{1}{2T} \int_{-T}^{+T} n(t) \, n(t + \tau) \, dt$$

$$= \frac{A^2}{2} \cos \omega\tau + \phi_n(\tau) \quad .. \quad .. \quad .. \quad (13)$$

Thus the correlation consists of two parts; one due entirely to the signal and the other due to noise. The signal component is periodic with the same frequency as the signal thus enabling the signal to be detected. The noise component rapidly decreases to zero since the noise is incoherent with itself for finite values of τ.

To illustrate this, let us consider a simple numerical example. Let the received signal, which we suppose to be sampled, be

$8\ 13\ \bar{4}\ 0\ \bar{8}\ 4\ 13\ 4\ 3\ \overline{12}\ \bar{3}\ 2\ 0\ 11\ 0\ \bar{9}\ \bar{6}\ 2\ \bar{4}\ 5\ \bar{2}\ 2\ \bar{1}\ \bar{7}\ 7\ \bar{1}\ 8\ \overline{13}\ \bar{1}\ \bar{5}\ 7\ 11\ 4\ \bar{6}\ \bar{2}\ 1\ 0\ 7\ \bar{2}$ where the bar above a digit indicates a negative sign.

We wish to determine whether there is a periodic signal present in this received message and, if so, the frequency of the signal. In this simple example we must limit the duration of τ, and obviously cannot integrate over an infinite range,

We choose T to be equivalent to 15 samples, then multiply these 15 numbers by the 15 numbers displaced an integral number of digits. That is, we multiply the 15 numbers beginning with 8 by the 15 numbers beginning with 8 then 13 then $\bar{4}$ and so on. The 15 products so obtained for 24 consecutive displacements can be tabulated, each row of the table is then summed (being an approximation to integration) and the resulting sums are divided by $2T$ (i.e., 30).

A plot of $\phi_{11}(\tau)$ against τ is shown in Fig. 3 where a periodic variation can be seen of frequency six digits. We conclude, therefore, that the original signal contained a periodic signal of frequency six digits.

The mean amplitude peak of the auto-correlogram is approximately 13. Therefore, from Equ. (13) we have $A^2/2 = 13$ or $A = \pm 5$ units. In fact, the waveform assumed before noise was added was

$5\ 5\ 5\ \bar{5}\ \bar{5}\ \bar{5}\ 5\ 5\ 5\ \bar{5}\ \bar{5}\ \bar{5}\ 5\ 5\ 5\ \bar{5}\ \bar{5}\ \bar{5}\ 5\ 5\ 5\ \bar{5}\ \bar{5}\ \bar{5}\ 5\ 5\ 5$, etc.

Due to the fact that only a short duration was considered, and integration was not extended over all t, only a small amount of noise could be added in order to illustrate the method.

In practice, with an electronic correlator, say, when duration and integration can be large, signals can be accurately detected in noise of considerable power.

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NEW YEAR HONOURS

Knights Bachelor

Gerald Clayton Beadle (Director, Television Broadcasting, B.B.C.).

Cecil Dannatt (Vice-Chairman, A.E.I. Ltd.).

Edward Robert Lewis (Chairman, Decca Navigator Co. Ltd., and Decca Radar Ltd.).

Alfred Charles Bernard Lovell (Professor of Radio Astronomy, University of Manchester and Director, Jodrell Bank Experimental Station).

C.B.

William Henry Stephens (Director-General of Ballistic Missiles, Ministry of Aviation).

C.B.E.

F. E. McGinnety (Director-General of Inspection, Ministry of Aviation).

A. J. Young (Managing Director, English Electric Valve Co. Ltd.).

O.B.E.

H. W. Baker (Supt. Engineer, London Television Studios, B.B.C.).

P. H. Dimmock (Head of Television Outside Braodcasts, B.B.C.).

S. J. Giffen (Senior Assistant Telecommunications Controller, Northern Ireland, G.P.O.).

M.B.E.

R. B. D. Dickinson (Executive Engineer, G.P.O.).

F. J. G. Haines (Engineer-in-Charge [sound], Cardiff, B.B.C.).

J. F. Lucas (Principal Station Radio Officer, Government Communications H.Q.).

J. W. Murray (Chief Engineer, Nigerian Broadcasting Corpn.).

B.E.M.

O. C. Baldock (Radio Operator, Government Communications H.Q.).

L. N. Sparks (Technical Officer, Post Office Research Station, Dollis Hill).

Transistor Measurements

POWER-GAIN AND NOISE FACTOR AT FREQUENCIES UP TO 100Mc/s

By B. N. Harden, M.Sc., A.M.I.E.E.,* and R. W. Smith, B.Sc.*

SUMMARY. *Equipment for the direct measurement, at frequencies from 0.5 to 100 Mc/s, of the power gain and noise factor of transistors operating as small-signal common-emitter amplifiers is described. The results of a few typical measurements are included, made under defined conditions of external feedback.*

Although the selection of a specific type of transistor for an amplifier depends on a variety of factors, a knowledge of power-gain and noise-factor, especially under optimum or near-optimum conditions, is useful in assessing transistor performance. The power gain of an amplifier stage is often estimated from values of the two-port parameters or from equivalent circuits but measurements have shown that appreciable discrepancies can occur between values calculated in this way and directly-measured results. The equipment described here has been in use over a long period for the direct determination of both power-gain and noise-factor performance of transistors operating under small-signal conditions in the common-emitter configuration.

The same units are used for both measurements at a number of frequencies in the range 0·5 to 100 Mc/s and the methods may be easily modified to suit specific requirements and available auxiliary equipment.

The basic methods of measurement are reviewed and some details of the circuits given, followed by a few typical results.

Definition of Measured Parameters

Power Gain

The maximum available gain,

$$\left(\frac{\text{power to load}}{\text{power available from the source}}\right)$$

and conditions for stability, have been derived in various ways[1,2,3] and may be expressed by equations in terms of two-port parameters of the amplifier stage. To define the conditions of measurement used here, these equations are briefly discussed using the H parameters of Fig. 1 from which they may conveniently be derived.†

Maximum power gain (G_{max}) is obtained for conjugate matching at both input and output terminals; i.e., when $Y_L = Y\dagger_{out}$ and $Z_s = Z\dagger_{in}$, giving

$$G_{max} = \frac{|H_f|^2}{2\,\mathrm{Re}\,(H_i)\,\mathrm{Re}\,(H_o) - \mathrm{Re}\,(H_f H_r) + [\{2\,\mathrm{Re}\,(H_i)\,\mathrm{Re}\,(H_o) - \mathrm{Re}\,(H_f H_r)\}^2 - |H_f H_r|^2]^{\frac{1}{2}}} \quad .. \quad (1)$$

with the unconditional relations,

$$\left.\begin{aligned}\mathrm{Re}\,(H_i) &> 0\\ \mathrm{Re}\,(H_o) &> 0\end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad .. \quad 2(a)$$

and $2\,\mathrm{Re}\,(H_i)\,\mathrm{Re}\,(H_o) - \mathrm{Re}\,(H_f H_r) > |H_f H_r| \quad .. \quad 2(b)$

In these expressions Re denotes the real part.

The stability conditions may be satisfied at some frequencies by the parameters of the transistor alone but, in other cases, external feedback must be applied

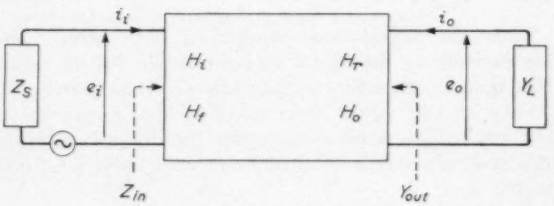

Fig. 1. *Equivalent circuit of amplifier*

so that the overall stage parameters satisfy Equ. (2). Three common practical cases arise:

(i) the stage is stable at any frequency for which $H_r = 0$. Suitable external feedback is applied to neutralize completely both real and imaginary components of the internal feedback (h_r) of the transistor; i.e., $H_r = 0$. (Lower-case letters are used for the transistor parameters and capitals for the overall-stage parameters.) This is the case of 'unilateralization' and Equ. (1) reduces to

$$G_{max} = \frac{|H_f|^2}{4\,\mathrm{Re}\,(H_i)\,\mathrm{Re}\,(H_o)}$$

(ii) the stage may be stable if the internal feedback of the transistor is only partially neutralized by external feedback. A common practical case is that of 'lossless neutralization' in which only the imaginary component of h_r is cancelled; i.e., $\mathrm{Im}\,(H_r) = 0$

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† The H parameters are defined by the following relationships between the terminal voltages and currents: $e_i = H_i i_i + H_r e_o \quad i_o = H_f i_i + H_o e_o$.

(iii) the transistor parameters satisfy Equ. (2) so that no external feedback is required for stability.

The H parameters above may be replaced by the corresponding elements of the Z and Y two-port parameters.

In common-emitter amplifier stages, external feedback is conveniently applied by a parallel network such as that in Fig. 2. This arrangement is used in the test equipment so that the corresponding forms of Equ. (1) and (2) expressed in terms of the overall Y parameters of Fig. 2, are applicable. The three cases above then become

(i) 'Y-unilateralization'

Here $ny_F = y_r$ (n is the turns ratio)

i.e., $ng_F = g_r$ and $nb_F = b_r$

and the power gain G_u is given by

$$G_u = \frac{|y_f - y_r|^2}{4\left(g_i + \frac{g_r}{n}\right)(g_o + n\,g_r)} \qquad .. \quad .. \quad (3)$$

Only the product $n\,y_F$ is unique so that by a suitable choice of n an optimum value of G_u is obtained. This occurs with negative values of g_r for $n = -\sqrt{\frac{g_o}{g_i}}$ giving

$$G_{uo} = \frac{|y_f - y_r|^2}{4(\sqrt{g_i\,g_o} - g_r)^2} \qquad .. \quad .. \quad .. \quad (4)$$

(ii) 'Y-lossless neutralization'

$n\,y_F = j\,b_r$; i.e., $g_F = 0$ and $n\,b_F = b_r$

giving

$$G_n = \frac{|y_f - j\,b_r|^2}{K + [K^2 - g_r{}^2\,|y_f - j\,b_r|^2]^{\frac{1}{2}}} \qquad .. \quad (5)$$

where $K = 2\,g_i\,g_o - g_r\,g_f$

(iii) No external feedback

$y_F = 0$ giving

$$G = \frac{|y_f|^2}{(K + b_f\,b_r) + [(K + b_f\,b_r)^2 - |y_f y_r|^2]^{\frac{1}{2}}} \qquad (6)$$

It is therefore in accordance with the definitions of Equs. (3), (5) and (6) that the three conditions of 'Y-unilateralization', 'Y-lossless neutralization' and 'No external feedback' will refer in the remainder of the paper. The definition of Equ. (4) is not used since the value of n for optimum G_u is a function of individual transistor parameters; when working with a wide range of transistor types it has been found more convenient to choose a suitable value of n at any frequency and use the near-optimum conditions of Equ. (3). In general, the differences are small. For Y-lossless neutralization, Equ. (5) shows that G_n is independent of the feedback turns ratio.

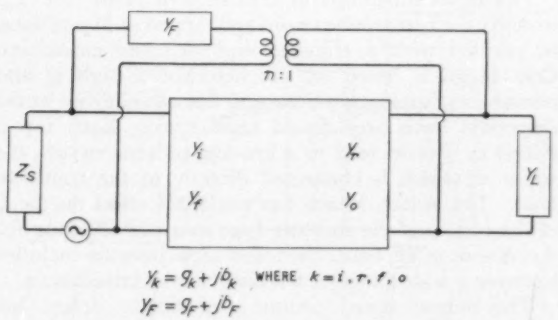

Fig. 2. *Equivalent circuit of neutralized amplifier*

All the equations may be simplified at particular frequencies and operating conditions as the y-parameters vary. For instance, at some frequencies, y_r may be small enough to be ignored, so that Equs. (3), (4), (5) and (6) all reduce to

$$G = \frac{|y_f|^2}{4\,g_i\,g_o}$$

Noise Factor

For a transistor amplifier, the noise factor

$$\left(\frac{\text{total output noise power}}{\text{output noise power thermally generated by the source}}\right)$$

is a function of the operating conditions (e.g., emitter current, source impedance, etc.) and a minimum noise factor may be obtained by suitable adjustment of the circuit. The equipment described here is directly suitable for noise-factor measurement under the conditions of the preceding section (Power Gain). Values obtained are, thus, not necessarily the minimum since operating conditions for maximum power gain are, in general, not the same as those for minimum noise factor.[4] However, since in practice a compromise between the two requirements may have to be adopted (especially at the higher frequencies where the noise of succeeding stages becomes important if the first stage gain is too low) the values measured under the above conditions have practical value.

Methods of Measurement

The equipment was originally developed to cover the frequency range 0·5 to 25 Mc/s and later another unit was added for the region about 100 Mc/s. Two different basic methods have thus been used but, in each case, there is a single common-emitter test stage with provision for conjugate matching at both input and output and a parallel external feedback network.

For frequencies from 0·5 to 25 Mc/s, power gain is measured by the method of Holmes et al[5] and a circuit diagram is given in Fig. 3. The transistor input is conjugately matched by a tapped coil (L_1) in parallel with a variable capacitor (C_1) to a signal generator and resistor (R_1) of total resistance R. The transistor output is matched by another tapped coil (L_2) and capacitor (C_2) to a load resistor (R_2) of value $4R$, and the feedback network (C_F, R_F) is adjusted for the required neutralization conditions defined in the preceding section (Power-Gain). If v_1 is the open-circuit voltage of the signal generator, the maximum available power to the transistor network is $\frac{v_1{}^2}{4R}$ and if v_2 is the voltage across the load resistor, the output power is $\frac{v_2{}^2}{4R}$. The power gain of the stage is thus given by $\frac{v_2{}^2}{v_1{}^2}$ or in decibels by

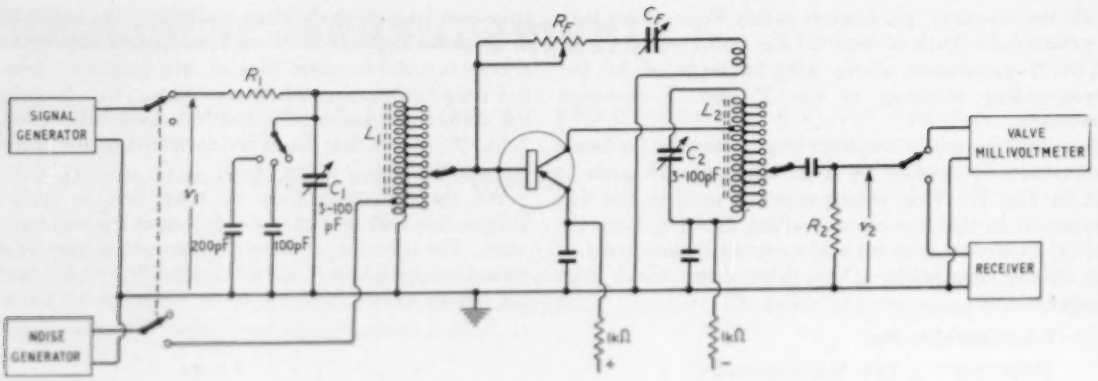

Fig. 3. *Units for frequency range* 0.5—25 *Mc/s*

$20 \log_{10} \frac{v_2}{v_1}$. The voltage v_2 is conveniently measured by a valve millivoltmeter which may be directly calibrated by the signal generator. Allowance is made for coil losses to obtain the power gain of the transistor.

With the same settings of the matching and feedback controls, noise factor is measured by replacing the signal generator and resistor R_1 by a diode noise generator,* connected to a suitable tap on the input tuned circuit so that the transistor is now matched to the noise generator, and then following the usual procedure of setting the noise-diode current to give double the output noise-power obtained with zero diode current. The change in output noise power may conveniently be observed on a calibrated millivoltmeter or by using an accurate 3-dB attenuator in the i.f. section of a low-noise receiver.

For the 100-Mc/s region, as shown in Fig. 4, the transistor input is matched directly to the output resistance of a signal generator and the transistor output matched to a stepped attenuator followed by a receiver. Nominal impedances are 75 Ω. The power gain, after allowance for any coil losses, is given by the change in attenuator setting required for constant receiver output when the signal-generator output is transferred from the input of the test stage to the input of the attenuator. Alternatively, a fixed attenuating pad may precede the receiver and adjustments be made on the output attenuator of the signal generator.

For measurement of noise factor, a diode noise generator replaces the signal generator and the noise output is observed by a low-noise receiver with a 3-dB attenuator in the i.f. section: When the stage gain is low, the receiver noise must be taken into account by using the relation for two cascaded stages,

$$F_M = F_T + \frac{F_R - 1}{G_T}$$

where F_M is the measured overall-noise factor,

F_T is the test-stage noise factor,

F_R is the receiver-noise factor, and

G_T is the test-stage power gain.

* Suitable generators using temperature-limited noise diodes and having output impedances in the range 50–70 Ω are available commercially.

If the correction term $\frac{F_R - 1}{G_T}$ is large, for example when a low-noise receiver is not available, improved accuracy may be obtained by inserting in front of the receiver a second transistor stage operating under the same conditions as the test stage and using the relation,

$$F_M = F_T + \frac{F_T - 1}{G_T} + \frac{F_R - 1}{G_T{}^2}$$

Details of Circuits and Operation

0·5 — 25-*Mc/s Units*

Six units, each fitted in a small die-cast box 7 in. × 4½ in. × 2 in., operate at frequencies near 0·5, 1, 3, 6, 13 and 25 Mc/s, the only differences in construction being in the component values. A unit covering the complete frequency range could be made with plug-in and switched components but separate small units have advantages in simplification of layout and wiring, with consequent reduction of unwanted couplings.

Nominal values of input matching resistors (R_1) range from 2·5 kΩ to 560 Ω at 0·5 and 25 Mc/s respectively, output resistors (R_2) giving the required ratio, being selected by d.c. measurements after allowance for the signal-generator output resistance. These values may easily be changed to permit matching to transistors with impedances outside the normal range.

The input tuned circuit consists of a small coil (L_1) wound on a Ferroxcube toroid and tapped in 11 positions, in parallel with a trimmer and switched capacitors. Coil design is based on the need for a high Q and convenient tapping positions and the values given in the Appendix have been found satisfactory. Each tap is joined by a short lead to a low-loss ceramic switch, the wiper of which is connected directly to the transistor base. The switch, which has negligible effect on the Q of the coil, is of the shorting type to avoid breaking the d.c. circuit to the base. Switched capacitors are included to cover a wide range of transistor input impedances.

The output tuned circuit is of similar design but without the switched capacitors. Coil inductance is higher to give the higher dynamic resistance required to

minimize coil loss. The collector is connected to a convenient coil tap chosen to suit most transistors and only in the case of very high output resistance does this tap require alteration.

The feedback network comprises a phase-reversing winding of about two to three turns wound over the tuning coil, in series with a variable capacitor (C_F) and a miniature carbon potentiometer (R_F). Bias circuits are decoupled and a screen fixed between the input and output circuits to prevent stray coupling. The transistor holder is of small brass tubes arranged so that most types of transistor fit easily into them and adaptors are used where necessary. A screen is provided to enclose the transistor completely. For transistors with high gain at the low frequencies, any possible damage from transients arising when the input taps are switched may be averted by detuning the output circuit before each change.

100-*Mc/s Unit*

The 100-Mc/s unit provides continuously-variable matching at input and output, the inductances being single-turn loops each rigidly mounted on a Perspex support, with sliding taps, designed to reduce series inductance, pivoted at the centres of the loops. Output connections are made by spring-loaded plunger-type contacts in rotating cups at the pivots. The input signal is tapped into the input loop and the collector connected to the non-earthy end of the output loop. A separate smaller feedback loop is mounted parallel to the output loop. Both input and output loop assemblies are separately screened and positioned so that the transistor holder is located in the space between them with minimum lead lengths. Additional screening is provided between the base and collector leads.

Neutralization

When external feedback is applied, measurement procedure with both types of unit involves successive adjustments of matching controls and feedback network. The matching controls are first set for maximum output. A signal is then fed into the transistor-output circuit with a sensitive receiver connected to the transistor-input circuit, care being taken that the matching conditions are not altered in this process. The feedback controls are then set for minimum output on the receiver, both R_F and C_F being adjusted for 'Y-unilateralization' and R_F set to zero for 'Y-lossless neutralization'. The new feedback conditions result in changed terminal impedances of the transistor network so that the matching controls need to be readjusted. The procedure is repeated as necessary. With experience of any specific type of transistor, successive adjustments may be kept to a minimum.

Accurate setting of the feedback controls is required for stability and for consistent results with transistors of high gain and output resistance.

Accuracy of Measurement

The accuracy depends on the quality of the available auxiliary equipment; for example, on the attenuators, millivoltmeter calibration, etc., but is largely controlled by the accuracy of estimation of coil losses. To determine these coil losses, the two tuned circuits for each unit were taken separately and fixed resistors substituted for the transistor terminal resistance. After tuning and matching, the difference between the maximum available power and the measured output power was calculated and represents the power loss in the coil for the appropriate tap position. By using a series of resistance values, a curve of power loss against tap position was obtained for each unit. For very high values of output resistance the above method is inaccurate and coil loss was calculated from the ratio of unloaded to loaded Q. Good agreement between the two methods was obtained in the overlapping region.

The estimated coil losses for the 0·5 to 25-Mc/s units are 0·5 $\pm$ 0·25 dB for the input coils and range up to 2·5 $\pm$ 1 dB for the output coils. Estimated coil losses in the 100-Mc/s unit are of the order of 0·5 dB.

Typical Results

To show the type of results obtained with the equipment, some measured values, corrected for coil losses, are given. They are confined to a few randomly-chosen samples, mainly of commercially-available v.h.f. transistors giving useful gain at 100 Mc/s, and the examples have been selected primarily to illustrate some of the variations of the parameters occurring with changes in frequency and operating conditions. They should be

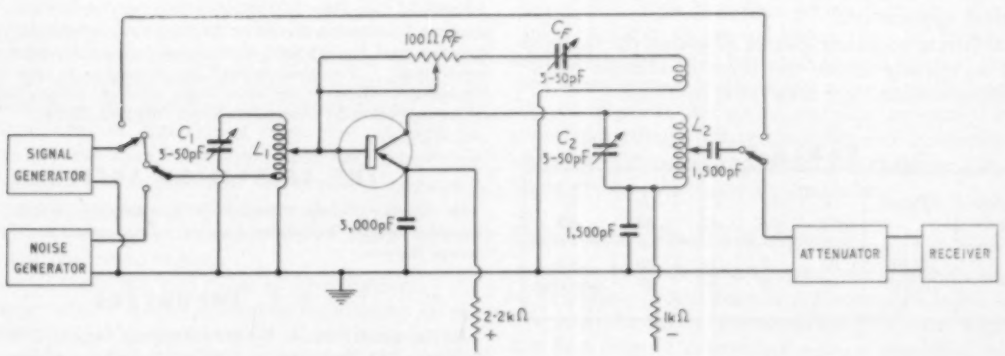

Fig. 4. *Unit for* 100-*Mc/s region*

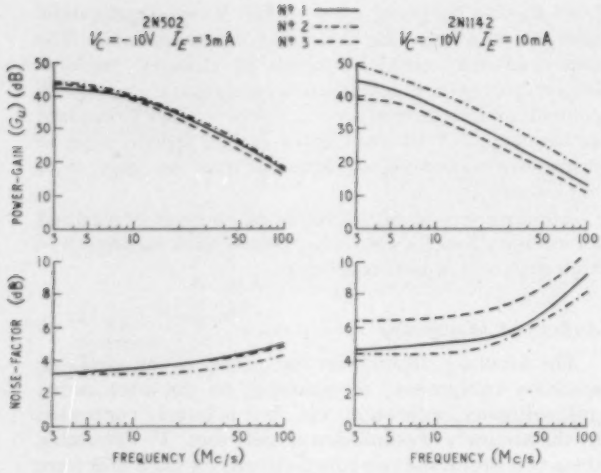

Fig. 5. *Y-unilaterized power gain and noise factor*

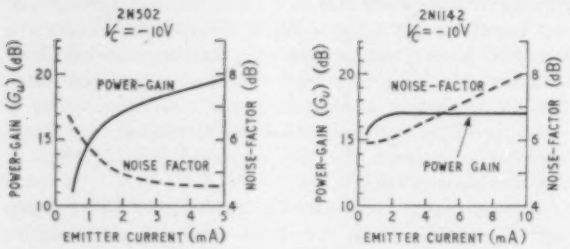

Fig. 6. *Variation of power gain and noise factor with emitter current; frequency* 100 *Mc/s*

regarded only as typical results and do not indicate the mean values or the spread obtained from large numbers of the same type of transistor.

Fig. 5 shows for frequencies from 3 to 100 Mc/s, the variations in power-gain and noise-factor for three samples of each of two types of 'Y-unilateralized' transistors. Typical changes observed when the emitter current was varied are given in Fig. 6 for two of the transistors at 100 Mc/s. Combination of results such as those of Figs. 5 and 6 assists in selecting operating points for practical applications.

Typical effects on power gain of adjusting the external feedback to operate under the three conditions in the section "Power-Gain" are illustrated in Table 1.

TABLE 1

Frequency Mc/s	Type	$-V_c$ volts	I_E mA	G_u dB	Gn dB	G dB
100	2N502	10	3	18	18	15
3	0C170	6	2	44	40	unstable

With some types of transistors, 'Y-unilateralized' power-gains of over 50 dB with corresponding noise factors no greater than 2 dB have been measured in the frequency range 0·5–1 Mc/s.

Conclusion

The equipment has proved satisfactory for the direct measurement of both power gain and noise factor of a wide range of transistors operating in the common-emitter configuration. The actual methods of measurement are easily adapted to meet any specific requirements arising from operating conditions or types of available auxiliary equipment.

Acknowledgements

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APPENDIX

Coil Data

Frequency Mc/s	Core Type No.	Input Coil (L_1) No. of Turns	Output Coil (L_2) No. of Turns	Collector Tap No. of Turns
0.5	FX 1231 B2	91	110	110
1	FX 1231 B2	60	76	50
3	FX 1596 B4	14	24	17
6	FX 1597 B4	17	24	13
13	FX 1598 B5	13	18	9
25	FX 1598 B5	12	12	4

FURZEHILL TO PRODUCE 'STROBOLYSERS'

Furzehill Laboratories announce that they have taken over the production, sales and service of the range of stroboscopes previously manufactured by Watford Instruments under the trade name of Strobolysers. Furzehill intend to introduce a new range of Strobolysers while, at the same time, offering service facilities for earlier models at 57 Clarendon Road, Watford, Herts.

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Sir Harold Bishop, director of Engineering, B.B.C., becomes president of the Television Society in succession to the late Sir George Barnes.

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Speed Control of D.C. Motors

TRANSISTOR EQUIPMENT FOR AUTOMATIC REGULATION

By M. J. Miller* and G. V. Buckley, A.M.I.E.E.*

Automatic speed control of small d.c. motors is usually obtained by electromechanical governors. These governors add little to the bulk of the motor and produce, for many purposes, a satisfactory speed performance. However, where a high degree of speed constancy is required, as in sound-recording systems, the performance provided by these governed motors does not compare with that of the a.c. synchronous motor. In addition, they have the disadvantage of producing governor-contact electrical interference and of usually requiring considerable maintenance in order to obtain consistently good results.

With the increasing development of servomechanisms, semiconductors and miniature components it has become possible to design speed-control systems for d.c. motors which are economic in price, physical size and current consumption, giving speed performances which are superior and more reliable than those obtained from electro-mechanically governed motors. In the field of sound recording, the speed performance obtained from small portable tape recorders fitted with servo-controlled d.c. motors can be better than that obtained from large studio machines using a.c. synchronous motors.

Motor Control Systems

Most small low-voltage d.c. motors of the shunt field or permanent-magnet field varieties are fitted with electro-mechanical governors in order to obtain as good a speed performance as possible. These governors are generally of the centrifugal type in which pairs of contacts opening and closing, insert or short out resistors connected in the armature supply circuit. They require considerable maintenance for consistent results and generate serious electrical interference due to the action of the contacts interrupting the armature current. Although these disadvantages can be largely overcome by using an electronic device, such as a transistor, in conjunction with the governor so that alteration of the armature current is transferred from the contacts to the transistor, the speed constancy attained is still not good for high-grade equipment. In order to approach or improve on the speed performance derived from a.c. synchronous motors it is necessary to use some type of servo-system with d.c. motors.

In general, there are two types of servo-systems,[1] the 'on-off' type which is the electronic equivalent of the electro-mechanical governor and the continuously-variable type, as defined by the diagram of Fig. 1. The second type gives a superior speed performance and is more flexible in the design of feedback loops. In this class, two designs are considered, one employing a phase-lock system and the second a simple velocity feedback system. In the first design, an a.c. signal of frequency proportional to the speed of the motor is phase-locked to a reference-frequency signal so that the long-term speed stability is that of the reference-frequency signal source. In the second design, an a.c. signal of frequency proportional to the speed of the motor is applied to a frequency-discriminating network which converts frequency-error to a control voltage so that the long-term speed stability is a function of the conversion gain of the system. In this second system, in contrast to the first, there is a permanent change of speed for a permanent change of motor load, the precise amount depending upon the value of the conversion gain. In both systems the short-term speed stability is dependent upon the characteristics of the control circuits and of the motor. An example of the use of these two systems is provided by some experiments

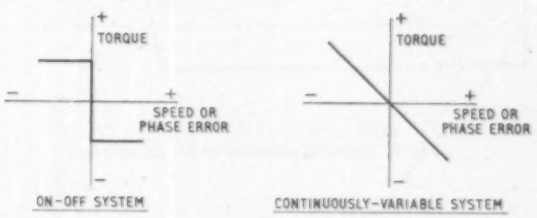

Fig. 1. *Types of servo-systems*

made with tape recorders where with the first system, speed stability figures of better than 0·1% were fairly easily obtained and, with the second system, the design problems became increasingly more difficult in attempting to improve the speed stability beyond 0·2%. In sound-recording systems the difference in the above two speed performances represents the difference between a fairly good and a first-rate machine.

Phase-Lock Feedback System

In this method an a.c. signal derived from the motor of frequency proportional to motor speed is fed back via two loops to the servo-amplifier. The a.c. signal from the first loop is compared with a standard reference-frequency a.c. signal and an error-signal proportional to phase difference is derived. The a.c. signal from the

* *Designs Dept., British Broadcasting Corporation*

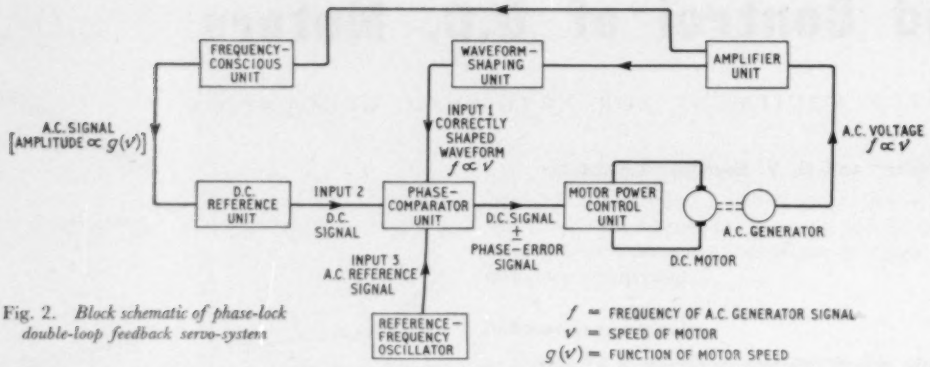

Fig. 2. *Block schematic of phase-lock double-loop feedback servo-system*

second loop controls the value of a stabilized d.c. voltage producing a d.c. signal which has a pre-determined voltage-frequency characteristic. The phase-error signal and this d.c. signal are combined and constitute the motor control voltage. The shape of the d.c. signal characteristic ensures that the motor starts automatically, and that it runs up to approximately its correct value without locking at some lower spurious speed. When

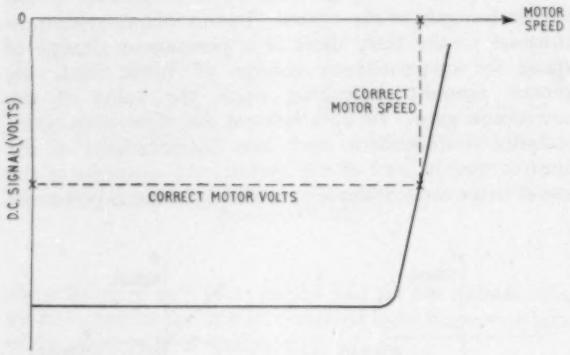

Fig. 3. *Output characteristic of d.c. reference unit*

the motor speed has reached its approximate correct value, the phase-error signal locks the motor speed or frequency to that of the standard reference-frequency a.c. signal.

This sytem is illustrated by the block schematic diagram of Fig. 2. An alternating voltage having a frequency proportional to the motor speed is derived from an a.c. generator which is coupled to the d.c. motor. The generator can take the form of a magnetic or light system but, for battery economy and reliability, perhaps the magnetic system is to be preferred using a combination of an ordinary mild-steel toothed wheel fixed on the motor shaft and a pick-up coil with a soft iron core mounted on the motor housing. By passing a direct current through the coil, the core is magnetized and if the toothed-wheel has n teeth and rotates at r revs/sec, a voltage of frequency, $f = nr$ (c/s) is generated in the coil. This voltage is amplified, if necessary, in the amplifier unit which provides two outputs for the two feedback loops, one via the waveform-shaping unit and the other via the frequency-conscious unit.

The first loop via the waveform-shaping unit, which shapes the signal to a desired form and rejects any spurious information, provides the phase-comparator unit with an a.c. signal whose frequency is proportional to motor speed, marked Input 1.

The second loop starts with the frequency-conscious unit which produces an a.c. output of amplitude which increases with motor speed. This signal is fed to the d.c. reference unit containing a stabilized source of d.c. voltage, from which a d.c. signal is obtained of value controlled by the a.c. signal. The output signal of the d.c. reference unit has a voltage-frequency characteristic shown in Fig. 3 and constitutes Input 2 to the phase-comparator unit.

The Input 3 to the phase-comparator unit is an a.c. reference signal of a specific frequency. In this unit, the phases of the two a.c. signals are compared and a phase-error signal is obtained of amplitude and sign depending on the phase difference between the two signals. The phase-error signal is superimposed on the d.c. signal (Input 2), adding to or subtracting from it and constitutes the output of the phase-comparator unit. This output voltage or motor control signal is fed to the motor power control unit which supplies the motor with power in such a manner that the speed of the motor is made synchronous with the frequency of the a.c. reference signal.

The practical application of this system can take a number of forms, one form using transistors being shown in Fig. 4. The phonic or toothed-wheel (W) and pickup or probe (P) are shown diagrammatically associated with the d.c. motor (M). The induced a.c. voltage from the probe is fed to the base of the amplifier transistor (V_1) which is biased by resistor (R_1) which also provides polarizing current for the probe coil. The output from this amplifier feeds the two loops via the resistor/capacitor combinations C_2 R_5 and C_3 R_4. In the first loop transistor V_4 acts as a clipper and feeds the clipped signal via the differentiating circuit (C_7 V_5 input resistance) to the quiescent pulse generator consisting of transistors V_5 and V_6. This section of the circuit is called the wave-shaping unit in Fig. 2. The pulse generator produces pulses only when triggered at the input to V_5 but the shape of the pulse is independent of trigger pulse shape. Therefore, the generator pulses

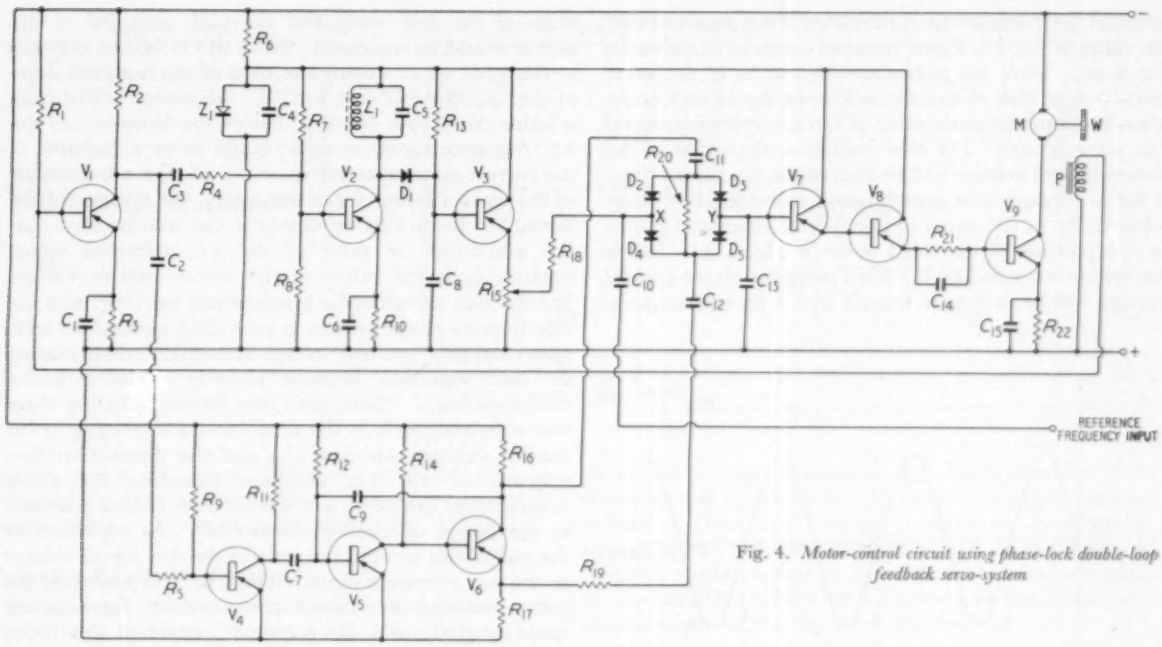

Fig. 4. *Motor-control circuit using phase-lock double-loop feedback servo-system*

carry the frequency and phase information of the motor a.c. signal but not any spurious information such as amplitude modulation of the a.c. signal. Two outputs are taken from the generator, one from the collector and the other from the emitter of V_6, the pulses occurring at the same instant of time and being of equal amplitude but opposite phase. These pulses are fed via capacitors C_{11} and C_{12} to the diagonal points of the bridge consisting of diodes D_2, D_3, D_4 and D_5, which acts as the phase comparator.

The second loop fed via C_3 R_4 starts with the frequency-conscious unit consisting of the circuit associated with the transistor V_2, which has an output characteristic determined by the LC collector load, that is, the a.c. output increases with motor speed or frequency. This output is fed to the diode D_1 which controls the emitter current V_3 by the value of the charge in capacitor C_8. When there is no a.c. input; that is, at zero motor speed, V_3 is bottomed and the emitter voltage is almost equal to that of the stabilized d.c. supply provided by the stabilizing circuit consisting of components resistor (R_6), zener diode (Z_1) and capacitor (C_4). This condition persists till the motor speed is very close to the correct speed when the diode D_1 starts to conduct, reducing the current in V_3 and hence the value of the emitter voltage. The curve connecting emitter voltage with motor speed is approximately of the form shown in Fig. 5 (a). To this d.c. voltage is added an a.c. reference voltage of a specific frequency; it could be of the same value as the frequency generated by the phonic wheel when the motor is running at its correct speed. The combined characteristic is shown in Fig. 5 (b) and constitutes the input voltage to the diode bridge (point X); this combined input corresponds with Inputs 2 and 3 of Fig. 2.

The diode-bridge or phase-comparator is a well-known configuration and it is known that when the pulses arrive via capacitors C_{11} and C_{12} all four diodes are made conductive and the voltage at point X is transferred to point Y charging up capacitor C_{13}. The pulses leave capacitors C_{11} and C_{12} charged, which hold the diodes cut-off till the next pair of pulses arrive. The

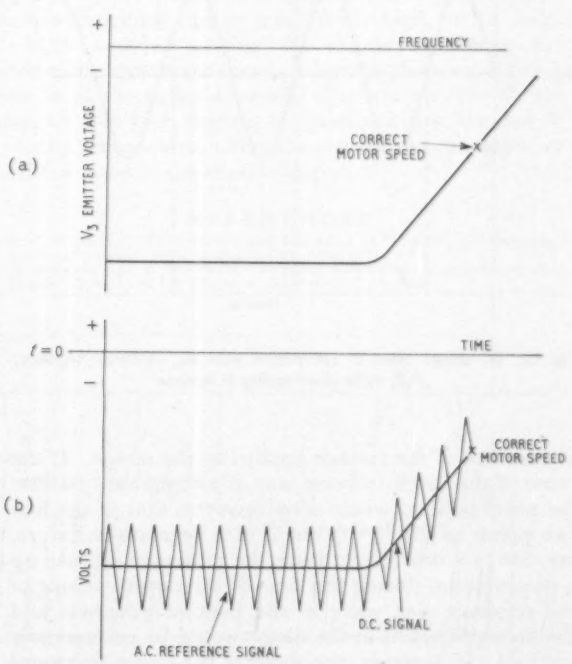

Fig. 5. (*a*) *Characteristic of V_3 emitter voltage;* (*b*) *characteristic of input voltage to phase comparator (point X)*

action of the diode bridge is simplified if it is assumed that the value of the d.c. signal remains constant as shown in Fig. 6 (a). Here the pulses are seen to be of the same frequency as that of the a.c. reference signal and occur when the instantaneous value of the a.c. reference-signal sine wave is zero. For this condition, the value of the motor-control voltage will be that of the d.c. signal only. If for any reason the motor speed increases, the phase relationship of the two a.c. signals will change as shown by the position of the third pulse in Fig. 6 (b). When the bridge is opened by the third pulse the motor-control voltage will be decreased from V to $V-\delta v$, which means

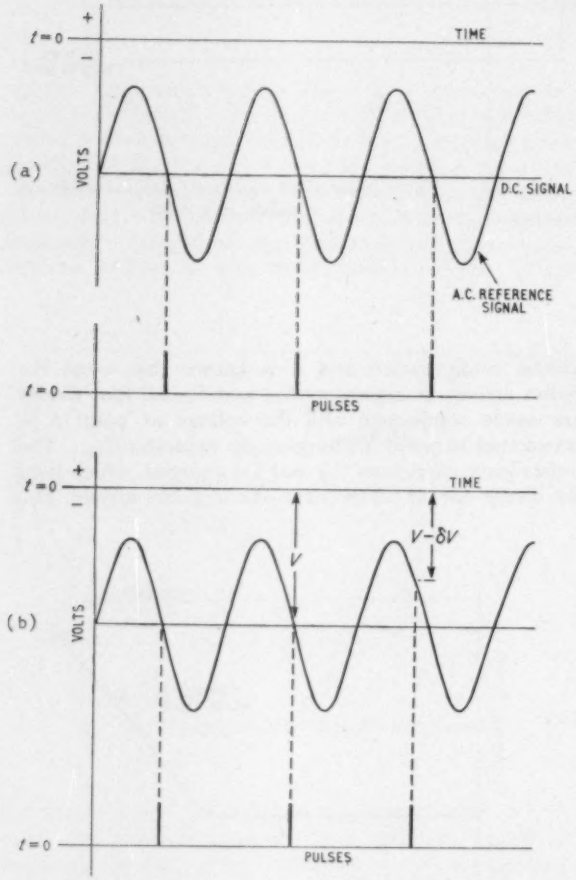

Fig. 6. *(a) Motor speed in synchronism with a.c. reference frequency; (b) motor speed tending to increase*

a reduction of the voltage applied to the motor. If the cause of the speed increase was of a temporary nature, the pulse position would soon revert to that of the first two pulses of Fig. 6 (b) but if of a permanent nature, say due to a decrease in load, the pulses would take up a new position during the time of the negative slope of the reference sine wave in the positive quadrant and the voltage applied to the motor would be permanently reduced. If, however, the speed of the motor decreased due to an increase of load the pulses would occur during the time of the negative quadrant of the negative slope of the sine wave and the volts supplied to the motor would be increased. From this it follows that the pulses must occur during the time of the negative slope of the a.c. reference sine wave for the system to maintain a stable phase-lock control. Hence the frequency of the a.c. reference signal must be equal to or a multiple of the correct motor speed frequency; if it is a sub-multiple of the correct motor speed frequency, the system will be unstable. From this argument it can also be seen that the amplitude or slope of the a.c. reference signal controls both the range of the motor control voltage (motor load variation for synchronous working) and the effectiveness of the system in providing good short-term speed stability (control voltage accurately compensating for load variation, without hunting or being under compensating). There are other factors affecting these characteristics such as the time constants relating to the control voltage capacitor C_{13} and the phase-frequency response of the d.c. amplifier following the phase comparator, but these are well-known factors common to the design of most servo-systems[2]. As explained at the end of this section, the ratio of the d.c. signal voltage to the a.c. reference-signal voltage is determined by the motor voltage-motor load characteristic for constant speed working and, by a proper choice of this ratio, automatic starting and avoidance of spurious speed running is ensured.

The motor control voltage developed across capacitor C_{13} is applied to a d.c. amplifier consisting of transistors V_7, V_8 and V_9, which provides a desired high-input impedance and the appropriate current gain for operating the final transistor, V_9, in whose collector the motor is connected. This system was used with a high-efficiency motor running at low speed (600 r.p.m.) and driving the capstan directly. In order to stabilize the system it was necessary to use the phasing circuits C_{14}-R_{21} and C_{15}-R_{22} and to connect the motor in the high-impedance collector circuit of transistor V_9.

The stability problem is eased considerably at the expense of overall efficiency if the value of total required torque is increased artificially so that the value of the percentage change of torque over the working range is reduced. This implies that the system now requires a smaller percentage change of control voltage, which eases both the feedback stability and spurious-speed locking problems. It also makes possible the operation of the motor in the emitter circuit of the final transistor without the use of phase-correcting networks, which simplifies the design of the d.c. amplifier. The spurious-speed locking problem can be explained by reference to Fig. 7 which shows the speed–volts characteristic of a motor and the envelope limits of the motor control voltage which consists of the a.c. reference signal superimposed upon the d.c. reference signal. If the minimum voltage value of this envelope is less than the voltage required by the motor to run at half speed then the motor will be locked to half speed, as shown in Fig. 7 (a). The limits of this envelope are defined by the control voltage variation required to maintain constant speed over a range of torque. Thus, if the limits of this envelope can be made smaller, which implies less required control voltage sensitivity, so as to avoid enclosing the half-speed voltage point, as shown in Fig. 7 (b), then the system

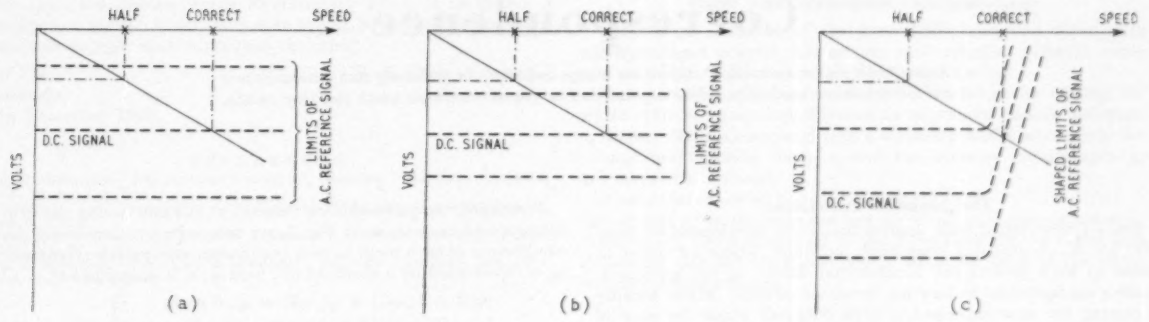

Fig. 7. *(a) Condition for spurious locking at half-speed; (b) condition for locking at correct speed; (c) shaped control-voltage characteristic to avoid spurious speed locking*

cannot lock at half speed. This, of course, can also be achieved by moving the whole envelope downwards as shown in Fig. 7 (c) which is the idea of Fig. 5 (b).

Velocity Feedback System

It was seen from the explanation of the circuit of Fig. 4 that the effect of the second loop was to bring the motor to a speed very close to its correct value, when the phase-lock effect took control locking the motor speed to the frequency of the a.c. reference signal. If such a high degree of speed accuracy is not required a very much simpler, smaller and cheaper unit can be designed as shown in Fig. 8 by using the principle of the second feedback loop of Fig. 4 by itself; that is, velocity feedback only. The output of the phonic wheel is fed to the base of the first transistor (V_1) via a variable resistor (P_2) which acts as an absolute speed adjuster. Tuned circuits in the collector and emitter of (V_1) provide the voltage–frequency characteristic at the collector of V_1 which, in conjunction with the diode (D_2), produce across capacitor (C_2) the correctly-shaped motor-control signal [similar to Figs. 3 and 5 (a)]. This signal is applied to the d.c. amplifier consisting of the three transistors (V_2, V_3 and V_4) and the zener diode (D_4), the d.c. motor being connected in the collector circuit of the output transistor (V_4). The zener diode (D_4) is necessary to transform the optimum-working point voltage as determined from the above frequency characteristic and which is developed across capacitor (C_2) to a very much lower value for the correct operation of transistor (V_4) without losing gain or system sensitivity. The diode (D_3) with its polarizing resistors (R_3 and R_4) form a protection circuit to limit starting surge current in transistor (V_4) and also limit the current in this transistor if the phonic-wheel signal fails for any reason.

The stabilized voltage produced by the components resistor (R_6), zener diode (D_1) and capacitor (C_1) ensures that motor speed is approximately independent of battery-supply variation. For instance, in a particular experiment it was found that for a battery voltage variation of 15—10 V the speed change was 0·3% using this circuit. This speed change was reduced to a negligible quantity by adding the circuit of the potentiometer (P_1) which alters the gain characteristic of transistor (V_1) with changing battery voltage. Thus, any change of speed is caused by a change of load and, for a given range of load, the improvement in speed change is determined by the loop gain of the amplifier which is a function of the frequency characteristic of the circuit of transistor V_1.

Conclusions

The two systems described are capable of providing a d.c. shunt field or permanent-magnet field motor with a speed performance equal to or better than that of an a.c. synchronous motor. In the case of the phase-lock system the motor speed is made synchronous with the frequency of an a.c. reference signal, and can be maintained over a given range of load and supply volts. It also offers the facility of varying the speed of the motor by altering the frequency of the a.c. reference signal and, as this frequency can be made fairly high, of the order 1–3 kc/s without making the stability problems too severe, the short-term speed variations (flutter and wow) can be made negligibly small. For systems that do not require such high degrees of speed stability, the simple velocity servo-system offers the advantages of simplicity and economy in space and cost.

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1 P. H. Hammond "The Analysis and Properties of ON-OFF and Saturating Servomechanisms", *RRE Journal* No 34, April 1954

2 H. Chestnut and R. W. Mayer, "Servomechanisms and Regulating System Design", Vols. I and II. Chapman and Hall Ltd., London.

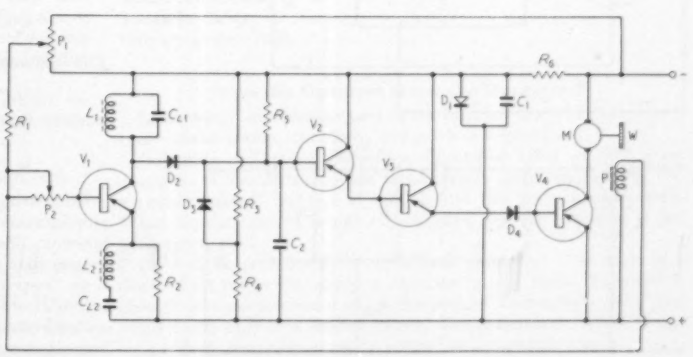


Fig. 8. *Motor control circuit using velocity-feedback system*

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

The Junction Transistor

SIR,—I have noted with interest the recent developments about Mr. M. K. Achuthan's phase-shift oscillator[1] where he seems to claim to have proved the fundamental nature of the grounded-emitter configuration in junction transistor circuits. While broadly accepting the view expressed in your editorial[2] about the need of a common standard configuration in terms of which circuit parameters are to be specified, I take the liberty of not accepting Mr. Achuthan's argument as a very convincing one. The fallacy in his circuit was clearly explained by Mr. Noordanus[3] and I do not see any reason why Mr. Achuthan still writes[4] so very vehemently against it. Let us, all the same, make ourselves more specific and re-draw the old phase-shift oscillator in both grounded-emitter and grounded-base configurations (after Mr. Achuthan) as shown in Fig. 1 (a) and Fig. 2 (a). Alongside are also drawn the respective amplifier circuits in Fig. 1 (b) and Fig. 2 (b). Since, however, Fig. 1 configuration clearly produces current gain there is no difficulty in understanding it, as is also the case with Mr. Achuthan. We therefore proceed straightaway with the grounded-base configuration as shown in Fig. 2.

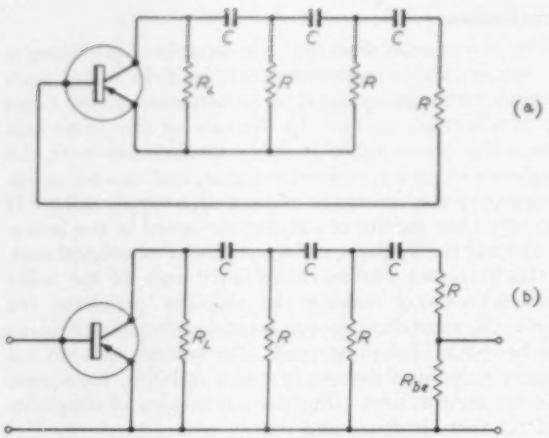


Fig. 1

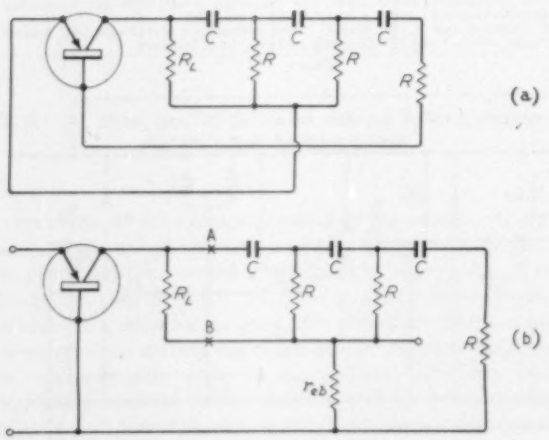


Fig. 2

Separating the phase-shifting network at AB and noting the loop voltages and currents as in Fig. 3, we can write by considering the equilibrium of each mesh in turn (equivalent emitter-base resistance r_{eb} is represented by a short-circuit, because it is insignificant),

$$\left.\begin{aligned} i_1(R + 1/j\omega C) + i_2(-R) + i_3 \,.\, 0 &= E_1 \,. \\ i_1(-R) + i_2(2R + 1/j\omega C) + i_3(-R) &= 0 \\ i_1 \,.\, 0 + i_2(-R) + i_3(2R + 1/j\omega C) &= 0 \end{aligned}\right\} \quad .. \quad (1)$$

If the network determinant is denoted by D, Equ. (1) yields,

$$i_1 = \frac{E_1}{D}\{(2R + 1/j\omega C)^2 - R^2\} \quad .. \quad .. \quad .. \quad (2)$$

$$\text{and } i_2 = \frac{E_1}{D} \,.\, R^2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (3)$$

$$\text{Also, } E_2 = i_2 \,.\, R = \frac{E_1}{D} \,.\, R^3 \quad .. \quad .. \quad .. \quad .. \quad (4)$$

$$\text{so that transfer function } T = \frac{E_2}{E_1} = \frac{R^3}{D} \quad .. \quad .. \quad (5)$$

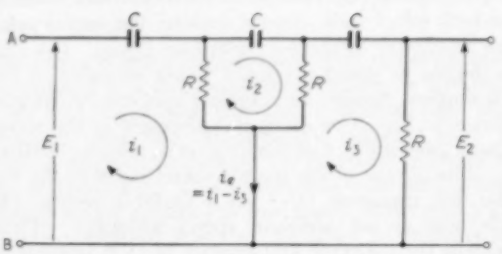


Fig. 3

Equ. (5) indicates that the network introduces a total phase-shift of π radians when the imaginary part of D vanishes.

By doing this, one finds from Equ. (1)

$$\omega_\pi = \frac{1}{CR\sqrt{6}} \quad .. \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

Thus, at $\omega = \omega_\pi$ Equ. (2) becomes,

$$i_1 = \frac{E_1}{D}\{3R^2 - 1/\omega^2C^2 + 4R/j\omega C\}$$

or, by inserting from Equ. (6),

$$i_1 = \frac{E_1}{D}\{-3R^2 - j4\sqrt{6}\,R^2\} \quad .. \quad .. \quad .. \quad (7)$$

The feedback current i_e is given from Fig. 3,

$$i_e = i_1 - i_2$$

or, from Equs. (3) and (7),

$$i_e = \frac{E_1}{D}\{-4R^2 - j4\sqrt{6}\,R^2\} \quad .. \quad .. \quad .. \quad (8)$$

Comparing Equ. (8) with (7), it is clear that there is current amplification at $\omega = \omega_\pi$, so that oscillation is possible. I would, however, like to point out here that it is the branch current that is greater than the collector or emitter (total) current and as such there is no "impossibility" involved as imagined by Mr. Achuthan[4]. To help him with a somewhat parallel example we may perhaps mention an antiresonant *LC* circuit where inductive and capacitive branch currents are each greater than the line or total current.

In conclusion, it may be observed that if a particular circuit oscillates in one line of argument, it must oscillate in another line

too, since the circuit cannot discriminate which of its points is strapped to earth. In fact it is only to our polarized thoughts that one form appeals more easily than the other.

Pye Ltd., R. R. Dutta Gupta.
Cambridge.
21st December 1960.

REFERENCES

[1] M. K. Achuthan, "The Junction Transistor", *Electronic Technology*, June 1960, Vol. 37, No. 6, p. 238.
[2] Editorial, "The Transistor Circuit", *Electronic Technology*, June 1960, Vol. 37, No. 6, p. 213.
[3] Ir. J. Noordanus, "The Junction Transistor", *Electronic Technology*", Sept. 1960, Vol. 37, No. 9, p. 361.
[4] M. K. Achuthan, "The Junction Transistor", *Electronic Technology*, Dec. 1960, Vol. 37, No. 12, p. 474.

Sir,—I should like to make some further comments on the question of the possibility of a fundamental mode of operation of the junction transistor, in reply to the paper[1] and the letter[2] of Mr. M. K. Achuthan and the letter of Mr. J. R. James.[3]

The purpose of this letter and an earlier one[4] is to stress the fact *that there is no basic transistor circuit.*

Mr. Achuthan's theory is based on the assumption that it is impossible to have a current or voltage amplification in a pssive *RC* network. But, if it is possible to devise an *RC* network (three-pole), which gives a 180° phase reversal between input and output current (or voltage), it is inevitable that the current in the third terminal of the circuit must be the sum of these two currents (Kirchhoff's law; see also reference 5).

Another argument in favour of a basic transistor circuit is given by Mr. J. R. James. He gives a method for determining this basic configuration by calculating the optimum oscillatory conditions for the three configurations. Only the earthed-emitter circuit gives an optimum which corresponds with experiments in practice.

The fallacy in his argument is in the fact that he uses a circuit determinant as a measure of the strength of oscillations. When we write down the circuit equations in the linear and sinusoidal case, we arrive at a determinant which must be zero to have undamped oscillations. Normally, this determinant will not be exactly zero and, with the right sign, there are also oscillations, but now they are growing in amplitude at a certain rate. A non-linear effect in the oscillator will limit the amplitude to a certain value.

In this case it is better to write down the general differential equations of the circuit. The solution of this set of equations gives the exponent a of e^{at} which describes the rate of growth of the oscillations.

For a circuit with linear elements, the exponent a must be the same for each *current* or voltage in the circuit. So an optimum in the rate of growth of the oscillations (optimum oscillatory condition) is *independent* of the way of considering the circuit.

In other words, the fundamental differential equations can change a bit in notation, depending on the choice of a certain current or voltage in the circuit (in which the differential equation is written), but the exponent a and the frequency of oscillation will always be the same.

Now, also in the case of the linear sinusoidal analysis, the system determinant depends on the way of analysing. Of course, all different determinants for earthed-base, earthed-collector, etc., muxt be zero for the same values of the circuit parameters and these are the values for which the exponent a of the analysis above is zero. This fact leaves the possibility that these determinants differ by a certain factor, which is a function of the value of the circuit components.

Therefore, these determinants are not directly representative for the strength of oscillation and conclusions derived from these determinants are not true.

Huizen (N.H.), Ir. J. Noordanus.
Holland,
1st January 1961.

REFERENCES

[1] M. K. Achuthan, "The Junction Transistor", *Electronic Technology*, June 1960, Vol. 37, No. 6, p. 238.
[2] M. K. Achuthan, letter to *Electronic Technology*, December 1960, Vol. 37, No. 12, p. 474.
[3] J. R. James, letter to *Electronic Technology*, September 1960, Vol. 37, No. 9, p. 361.
[4] Ir. J. Noordanus, letter to *Electronic Technology*, September 1960, Vol. 37, No. 9, p. 361.
[5] Thomas Roddam, "Cathode Follower Oscillator," *Wireless World*, March 1954, Vol. 60, No. 3, p. 106.

Valve and Transistor Configurations

Sir,—Your Editorial in the June 1960 issue and subsequent correspondence suggest that certain dual relations between valve and transistor configurations are worth mentioning.

Consider a simplified valve model (in low-frequency régime) for which only the voltage amplification factor μ and the anode resistance r_a matter. This is compared with a transistor model where only the current amplification factor α and the collector conductance g_c are taken into account.

If we define $\alpha' = \alpha/(1-\alpha)$; $g_c' = g_c/(1-\alpha)$
then α' and g_c' in the transistor are the 'dual' analogues of μ and r_a in the valve; this is illustrated by the *Y*-matrices (for the valve) and *Z*-matrices (for the transistor) given below for each of the three principal configurations.

Matrices For Valve And Transistor Configuration

Three Dual Pairs

Valve parameters: μ, r_a ($g_a = 1/r_a$; $g_m = \mu g_a$)

Transistor parameters: α', g'_c derived from α, g_c as follows: ($r'_c = 1/g'_c$; $r'_m = \alpha' r'_c$; $\alpha' = \alpha/(1-\alpha)$; $g'_c = g_c/(1-\alpha)$

1—Common-cathode triode

$$A = \begin{bmatrix} -\frac{1}{\mu} & -\frac{r_a}{\mu} \\ 0 & 0 \end{bmatrix} \qquad Y = \begin{bmatrix} 0 & 0 \\ -g_m & -g_a \end{bmatrix}$$

Common-emitter transistor

$$A = \begin{bmatrix} 0 & 0 \\ -\frac{g_c'}{\alpha'} & -\frac{1}{\alpha'} \end{bmatrix} \qquad Z = \begin{bmatrix} 0 & 0 \\ -r'_m & -r'_c \end{bmatrix}$$

2—Cathode-follower triode

$$A = \begin{bmatrix} \frac{1+\mu}{\mu} & \frac{r_a}{\mu} \\ 0 & 0 \end{bmatrix} \qquad Y = \begin{bmatrix} 0 & 0 \\ g_m & -(g_m+g_a) \end{bmatrix}$$

Common-base transistor

$$A = \begin{bmatrix} 0 & 0 \\ \frac{g_c}{\alpha'} & \frac{1+\alpha'}{\alpha'} \end{bmatrix} \qquad Z = \begin{bmatrix} 0 & 0 \\ r_m & -r_m+r_c \end{bmatrix}$$

3—Common-grid triode

$$A = \begin{bmatrix} \frac{1}{\mu+1} & \frac{r_a}{\mu+1} \\ 0 & 1 \end{bmatrix} \qquad Y = \begin{bmatrix} g_m+g_a & -g_a \\ g_m+g_a & -g_a \end{bmatrix}$$

Emitter-follower transistor

$$A = \begin{bmatrix} 1 & 0 \\ \frac{g'_c}{1+\alpha'} & \frac{1}{1+\alpha'} \end{bmatrix} \qquad Z = \begin{bmatrix} r'_m+r'_c & -r'_c \\ r'_m+r'_c & -r'_c \end{bmatrix}$$

Conventions: Input and output voltages are of the same sign. Input and output currents are in the same direction.

These relations may suggest more realistic models in which the above-mentioned duality is preserved by taking into account pairs of related quantities such as grid conductance and base resistance.

W. Proctor Wilson.
British Broadcasting Corporation,
Research Department,
Tadworth, Surrey.
16th December 1960.

"Can the Operand Stand the Operator?"

Sir,—May I attempt the part of the sceptical but—in Computor's eyes—praiseworthy schoolboy, and ask a few questions?

Computor (*Electronic Technology*, December 1960, p. 460) gives examples of operands that can stand certain operators fully, or "to a certain extent", or not at all. I find this very unsatisfactory. What are the tests by which one decides, and who or what is the authority?

Distance is presumably a directional quantity. −1 inch is a distance set off in the opposite direction to +1 inch. Potential is classed as non-directional so, according to Computor's rule, can stand minus only to a limited extent. But is that so? Where is the limit? Both quantities must be related to an arbitrary zero.

"−1 apple" may mean "take away 1 apple". It can also mean "add 1 negative apple", a negative apple being a body which,

when brought into contact with a positive apple, causes both to disappear. Computor declares there is no such thing as a negative apple. How does he know? Has he explored the entire universe? Until quite recently he would have been equally justified in saying there was no such thing as a negative electron. Did the rules of mathematics have to change when he was told about the positron?

What sort of 'bill of health' should one give electric charge? It is not directional or vectorial so, according to Computor, it ought to be unable to stand unqualified application of minus. Yet a charge equal to 6–8 units has a clear meaning. It would equally do so if authority had decided to call 'positive' what we conventionally call 'negative'.

Computor says "a negative resistance cannot exist at all unless a positive resistance of greater value exists elsewhere in the circuit" But there are circuits (containing, for example, a tunnel diode) where for a time at least the *net* resistance is negative, as shown by the exponential growth of oscillatory current. In principle, such a time could outlast the human race, and a schoolboy might happen to have come across no other kind of circuit, so naturally would regard negative resistance as a positive entity. In what respect would he be wrong?

The tests for examining operands for fitness seem to be somewhat hazy and subjective. I think the schoolboy would have more respect for mathematics if the rules could be seen to be permanent and objective, not depending on the current state of observation of the physical universe or on arbitrary conventions such as the sign of the electronic charge.

Bromley, * M. G. Scroggie.
Kent.
14th December 1960.

Sir,—We are here concerned with the use of mathematics as a tool for understanding actual phenomena. For a given problem, our first task is to make an effective model of the situation under discussion: this may be an equivalent circuit, but may also be a mathematical model or idea, which must be associated with an algebra appropriate to the situation.

For impedance theory the idea of a versor $x+jy$ is such a model, and the algebra of complex numbers is appropriate while, in electromagnetic theory, vectors are more suitable, with an algebra which includes vector and scalar products. For some circuit work, matrix algebra is required, and the fact that the products AB and BA are unequal fits in with the fact that exchanging two elements in, say, a ladder network, alters the effect of that network on the external world.

In any of these algebras we can split up a term in various ways, say, by writing $3R–2R$ for R or x^{n+1}/x^n for x, or by factorizing a given matrix A into $A_1 A_2 A_3 \ldots A_n$. But the terms obtained during such processes need not have any physical significance or realizability (unless they are final results). Thus in the expression $3R–2R$, the minus sign is a mere order to subtract $2R$ (which can be done) and no problem arises unless the term $-2R$ is divorced from its context. Such separation from context breeds the false notion that the term $-2R$ indicates the addition of a strange entity written in words as two "minus-R"; the separation often arises because a scale is used for which the zero is completely arbitrary and conventional. The Centigrade temperature scale, for instance, has an arbitrary zero whereas, with the Kelvin scale, negative temperatures cannot arise because the zero is at the temperature at which all molecular motion ceases. Likewise with potentials, we nearly always measure only relatively. The absolute zero of potential is that of a charge so far removed from other charges that further removal makes no difference, so that the charge is at infinity in the metrical sense, and has zero energy. Negative energy is a difficult concept.

Of all the laws of nature, perhaps the most firmly established is the second law of thermodynamics. It can be expressed in terms of information theory and in other ways, but in impedance theory briefly and succinctly by the statement Resistance is Positive. What is often called a 'negative resistance' is preferably described explicitly as a source of energy, however concealed. If negative resistances really existed, we could use them to replace generating stations and save coal, oil and uranium.

It could be argued that many of the contradictions of basic mathematics and of logic (such as the problem of the barber who shaves everyone in the village who does not shave himself) reduce to intense consideration of the question, "Can the Operand Stand the Operator?"

Computer.

21st December 1960.

New Books

Servomechanisms

By P. L. Taylor, M.A., A.M.I.E.E. Pp. 418 + xiv. Longmans, Green & Co. Ltd., 6 & 7 Clifford St., London, W.1. Price 42s.

This is an introductory book on closed-loop control. The book is in three parts; Part I introduces the idea of closed-loop control and gives a short account of data transmission and data computation. Part II develops the ideas of Part I and gives some physical descriptions of how servomechanisms operate and how they can be improved. The rest of the second part is devoted to describing the properties of the units used in servomechanisms such as amplifiers, phase-sensitive rectifiers and power drives of various sorts. The general mathematical theory of linear closed-loop systems is given in Part III which includes methods of practical analysis such as root-locus and harmonic response.

Students of servomechanisms must be bewildered by the number of text books existing on the subject, which cover very similar ground. It is not obvious that the material and treatment of the subject in this book has such particular advantages as to justify adding to the number. There does appear to be in fact some distinct drawbacks. The material for the early chapters appears to be drawn from rather a limited field of practice. For example, in the chapter on static amplifiers, one would have thought that by this time transistors would have at least deserved a mention. A chapter on digital and analogue computers is included, which is hardly directly relevant to the subject matter of the book. The treatment of the analysis of hydraulic drives on the other hand is so sparse as to be misleading. It is stated that compressibility effects are worse the greater the volume of oil. However, in the case of a jack, the natural frequency of the oil column is proportional to jack area. Thus, the problems due to oil compressibility are reduced by increasing the jack area and therefore the oil volume.

The author compares the root-locus method unfavourably with harmonic-response methods in numerical design. The account given of the root-locus method seems to overstress the labour involved in the construction of the diagrams. No mention is made of the simple mechanical aids, such as the Spirule or a protractor with a rotating cursor, which drastically reduce the time required to draw the root-loci. It is also felt that the utility of the root-locus method in the design and synthesis of compensation methods is not given proper weight. Thus, it is not every servo-engineer who would agree with the author's conclusions.

A.T.M.

R.S.G.B. Amateur Radio Call Book: 1961 Edition

Pp. 70. Published by the Radio Society of Great Britain, New Ruskin House, Little Russell St., London, W.C.1. Price 4s. (plus 6d. postage).

"Since the 1960 edition more than 450 new calls have been issued, more than 100 old calls have been re-issued, upwards of 1,000 changes of address have been recorded and about 250 calls have been cancelled. This edition reflects those additions and changes. It also lists Societies and Clubs in affiliation with the R.S.G.B."

Le Récepteurs de Télévision, Vol. 2

By W. T. Cocking, M.I.E.E. Translated by H. Piraux. Pp. 307 + vii. Dunod, 92 Rue Bonaparte, Paris 6e. Price 35 NF.

This book is a translation into French of the fourth edition of "Television Receiving Equipment", published by Iliffe Books Ltd.

Vol. 1, containing chapters 1–14, appeared some time ago and covers the television signal, the c.r. tube, focusing and deflection, and time-bases. Vol. 2 (chapters 15–30 and appendixes) treats video, i.f. and s.f. amplifiers, detectors, frequency-changers, sound-channel rejection, aerials, gain control, and e.h.t. and other power supplies.

Television Field Trials (of 405-line and 625-line systems in the u.h.f. and v.h.f. bands)
Pp. 152. Published by the British Broadcasting Corporation, 35 Marylebone High Street, London, W.1. Price 20s.

"In 1952 the Postmaster General set up a Television Advisory Committee to advise him on the development of television in the United Kingdom. The Committee in turn set up a Technical Sub-Committee and the present Report has been made by a Working Party of this Sub-Committee, which had the task of co-ordinating tests of television transmission in Band V (610–960 Mc/s). The Introduction to Part 1 of the Report explains the circumstances in which the tests were carried out and the Report itself gives the results of the field trials of 405-line television transmission in Band I (41–68 Mc/s) and Band V and also of 625-line transmissions in Band V.

The Report provides a record of a most comprehensive and valuable series of tests not previously undertaken in Europe, and it was decided, with the concurrence of all the organizations which participated in the work, the B.B.C. should publish the Report. . . ."

F.B.I. Register of British Manufacturers—1961 (33rd Edn.)
Pp. 1162. Published for the Federation of British Industries by Kelly's Directories Ltd. and Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 50s. post free.

"The F.B.I. Register, the comprehensive and accurate guide to a substantial cross-section of the British industry, contains lists of the products and services of nearly 8,000 member firms.

In addition to the classified buyers' guide there are seven other sections giving addresses of companies and firms, and information about Trade Associations, proprietary names, trade marks, etc. The French, German and Spanish glossaries give translations of every product term used in the main buyers' guide. . . ."

Electronic Equipment Reliability
By G. W. A. Dummer, M.B.E., M.I.E.E., Sen. Mem.I.R.E., and N. Griffin, A.M.Brit.I.R.E. Pp. 274 + xi. Sir Isaac Pitman & Sons Ltd., Parker Street, Kingsway, London, W.C.2. Price 45s.

". . . the primary aim of this book is to consider the many aspects of the subject of reliability and thus present a summary of the present knowledge in order to assist designers in obtaining maximum reliability in their equipment."

MEETINGS

I.E.E.

7*th February*. "Magnetic Properties of Thin Films for Computing Devices", by E. M. Bradley.

13*th February*. "The Road to Corporate Membership: What is the Responsible Experience Required?" Discussion to be opened by A. H. Mumford, O.B.E., B.Sc.(Eng.) and D. S. Rolfe, at 6 o'clock.

16*th February*. "Transistors and All That", Faraday Lecture by L. J. Davies, C.B.E., M.A., B.Sc., to be held at 6 o'clock at Central Hall, Westminster, London, S.W.1.

17*th February*. "A Comprehensive Introduction to Plasma Physics", by W. B. Thompson, B.A., M.A., Ph.D.

22*nd February*. "Coding Theory and its Applications", by J. M. Meggitt, M.A., Ph.D.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2 (except the Faraday Lecture on 16th February) and will commence at 5.30, except where otherwise stated.

Brit. I.R.E.

7*th February*. "Tunnel Diodes". Joint meeting with the Institute of Physics and Physical Society, at 2.15, at 47 Belgrave Square, London, S.W.1. (Admission by ticket only.)

8*th February*. "Perseus, a Medium-Scale Data Processing System", by G. Emery, M.A.

14*th February*. "Some Psycho-Acoustic and Engineering Aspects of Hearing Aid Design", by J. Jessop, M.Sc.

22*nd February*. "Transistorized Medical and Biological Electronic Equipment". Discussion.

These meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, at 5.30, with the exception of the meeting on 7th February.

The Television Society

10*th February*. "Transistorized Distribution Amplifiers", by B. Marsden.

23*rd February*. "Programme Problems and Possibilities Associated with 'Monitor'", by Huw Wheldon.

These meetings will be held at 7 o'clock at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.

The Society of Instrument Technology

28*th February*. "Three-Terminal Bridge Measurements and Automation", by V. S. Griffiths, Ph.D., B.Sc., to be held at 7 o'clock at Manson House, 26 Portland Place, London, W.1.

British Kinematograph Society

22*nd February*. "Carbon in Motion Picture Engineering and Electronics", by A. C. Conford, at 7.30 at the Central Office of Information, Hercules Road, Westminster Bridge Road, London, S.E.1.

The Society of Environmental Engineers

15*th February*. "The Development and Testing of Small Electronic Components", to be held at 6 o'clock at The Imperial College of Science and Technology, Prince Consort Road, London, S.W.7.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)
Deviations from nominal frequency for December 1960*

Date 1960 December	MSF 60 kc/s 1430–1530 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	NM	− 10
2	− 15·3	− 7
3	− 15·3	NM
4	− 15·3	NM
5	− 15·3	− 9
6	− 14·9	− 8
7	NM	− 6
8	− 14·8	− 5
9	− 14·5	− 5
10	− 14·4	NM
11	− 14·5	NM
12	− 14·2	− 4
13	− 14·1	− 2
14	− 14·3	− 2
15	− 13·9	0
16	− 14·6	− 1
17	− 14·9	NM
18	− 14·8	NM
19	− 14·7	+ 3
20	− 14·8	+ 2
21	− 14·7	+ 3
22	− 14·8	+ 5
23	− 14·7	+ 4
24	− 14·7	NM
25	− 14·6	NM
26	− 14·6	NM
27	− 14·3	NM
28	− 14·0	NM
29	− 14·3	+ 8
30	− 14·1	+ 9
31	− 14·3	NM

* *Nominal frequency corresponds to a value of 9 192 631 770 c/s for the caesium F,m (4,0)–F,m (3,0) transition at zero field. NM = No Measurement.*

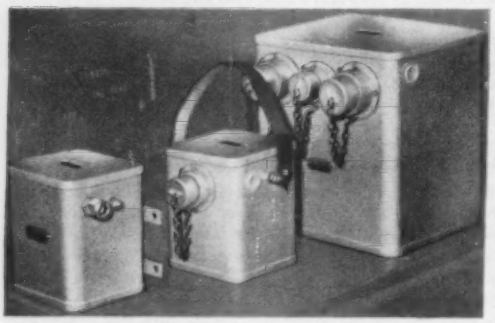
New Products

Low-Voltage Lighting Transformers

To meet a demand for safe working voltages in industry, Gresham Transformers have developed a range of low-voltage transformers complying with B.S.S.794.

Totally enclosed in robust steel cases, these transformers can be supplied for loads up to 1 kVA for floor mounting, wall fixing or as portable units. Units for loads up to 5 kVA are available where it is required to have one major low-voltage source.

The standard range of transformers have



the normal 200–250-V primaries and a choice of secondaries: 55–0–55 V or 25–0–25 V.

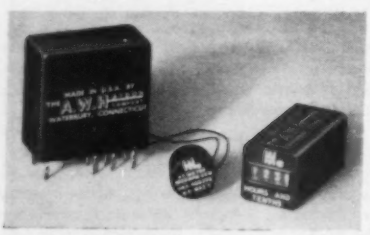
The Gresham Transformer Group,
Twickenham Road, Hanworth, Middx.

Micro-Miniature Motor and Timers

Two unusually small timing devices and an associated micro-miniature motor have recently been announced by the A. W. Haydon Co., U.S.A.

The 115-V 400-c/s single-phase hysteresis-type motor measures $\frac{3}{8}$ in. diameter by $\frac{9}{32}$ in. in length, and consumes less than $\frac{1}{2}$ W of power. The torque of a motor rated at $1{\cdot}5 \times 10^{-6}$ horsepower is 0·0005 ounce-inches, and its weight is given as $\frac{1}{3}$ oz.

The motor was specifically developed to



power a hermetically-sealed repeat-cycle timer measuring $\frac{1}{2}$ in. × 1 in. × 1 in. and weighing 1 oz. Three roller switches in the timer are driven at 1 r.p.m. by the motor, while the switches in the device are rated for 7 A (resistive) at 28 V d.c. and 5 A (resistive and inductive) at 115/230 V a.c. 60 c/s.

The second device powered by the motor is a miniature digital elapsed-time indicator. The four-digit indicator measures $\frac{1}{2}$ in. × $\frac{1}{2}$ in. × 1 $\frac{1}{16}$ in. long, and operates from a frequency of 400 c/s,

The A. W. Haydon Co.,
Waterbury, Connecticut, U.S.A.

Transistorized Phase Meter

A transistorized phase meter designed to measure the difference in phase angle between two signals in the audio and sub-audio frequency range has been introduced by Dawe Instruments.

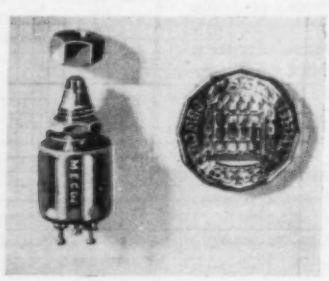
Known as the Type 630 transistor phase meter, it is suitable for signals in the frequency range 1 c/s to 25 kc/s with amplitudes between 0·03 V and 300 V. Two phase-angle ranges, 0–360° and 180°–0–180°, are provided and the instrument operates from the normal a.c.-mains supply or self-contained batteries.

Phase accuracy depends, to some extent, upon the relative amplitudes of the signals to be measured. With amplitude-balanced signals, on the 0–360° range the phase accuracy is ± 3° over the frequency range 10 c/s to 10 kc/s and ± 5° from 1 c/s to 10 c/s. With unbalanced signals, a typical error is ± 10° for a 3 : 1 or 1 : 3 unbalance from 10 c/s to 1 kc/s.

Dawe Instruments Ltd.,
Harlequin Avenue,
Great West Road, Brentford, Middx.

Miniature Rotary Potentiometer

The TP5 single-turn miniature potentiometer which is now in production by Miniature Electronic Components has been specially developed for the aircraft and guided-weapons industries. Compact in construction, it measures 0·5 in. diam. × 1·14 in. long.



This potentiometer is capable of dissipating $1\frac{1}{2}$ W at 80 °C; has a temperature range of −55 °C to 150 °C, and is available in ten standard resistance values from 50 Ω to 50 kΩ.

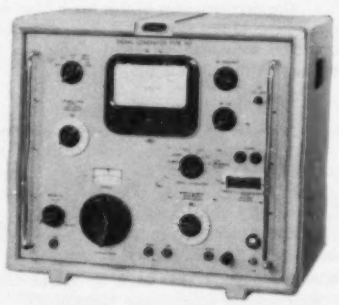
Miniature Electronic Components Ltd.,
The Lye,
St. Johns, Woking, Surrey.

Four-Phase V.L.F. Signal Generator

The new Airmec v.l.f. signal generator, type 257, operates over the frequency range 0·03 c/s to 30 c/s and provides four phase-related output signals simultaneously.

The 'Normal' output signal can be continuously varied from 500 μV to 50-V peak. The 'Reference' and 'Quadrature' outputs provide fixed-level (12 V) signals suitable for phase-measurement purposes, the 'Reference' being in phase with the 'Normal' output and the 'Quadrature' lagging by 90°. A fourth 'Variable-Phase' output provides a signal which is continuously adjustable in phase (with respect to the 'Normal' output) by means of a control on the front panel calibrated 0–360°. This signal is controllable also in amplitude over a range of 20 dB up to a maximum of 20-V peak.

The basic generator employs a motor-driven capacitor to modulate a high-frequency signal. The modulated signal is



then rectified and amplified to provide the very-low-frequency output.

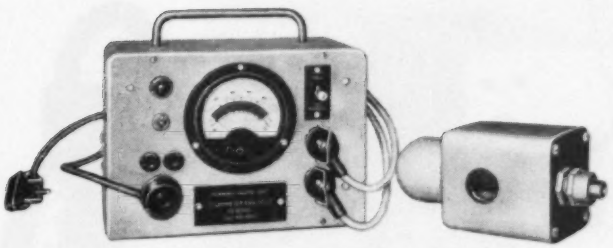
Airmec Ltd.,
High Wycombe, Bucks.

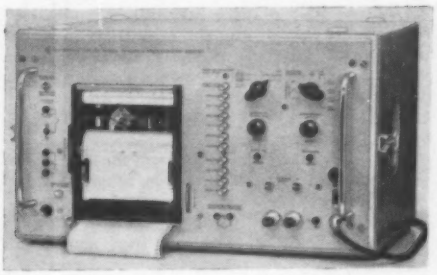
Cold-Cathode Ionization Gauge

A new high-vacuum gauge, either portable or panel-mounted, has been developed by Pulsometer Engineering. It is a Penning-type cold-cathode ionization gauge, which has a range 10^{-3} to 10^{-6} Torr.

The construction is a small ring anode with two-plate cathodes sealed in a glass bulb connected to the vacuum system. A permanent magnet produces a field along the axis of the electrodes.

Electrons leaving the cathode travel in helical paths towards the anode; the ionization current for a given electron current is,





therefore, very large and so can be measured directly on a microammeter without amplification.

Connection of the gauge head to the vacuum system is through a ½-in. vacuum union and to the control unit by a 4-ft length of high-voltage cable. The gauge head can be supplied with a flanged adaptor piece.

Pulsometer Engineering Co. Ltd.,
Nine Elms Iron Works,
Reading, Berks.

Motorized Turret Tuner

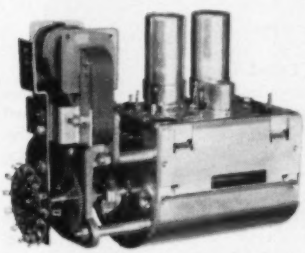
A motor-driven turret tuner designed to cover all television channels and f.m. channels has been introduced by Plessey. This unit, type PL14M, will facilitate the use of remote control and push-button techniques in domestic television receivers. No mechanical loading falls on push buttons, which are used only to switch electrical circuits.

Using the latest frame grid valves, the design of the turret affords accurate re-setting with minimum frequency drift; however, a fine tuner is available as an optional fitment,

The motor-drive unit uses a mains-voltage nduction motor. Accurate re-setting by self-indexing the turret, and automatic disconnection to prevent over-run when the motor is switched off, are provided by a special clutch.

Operating time of the tuner for 360° rotation is seven seconds.

The Plessey Co. Ltd.,
Ilford, Essex.



Digital Clock

A digital clock, type 354, produced by Microcell Electronics, has been primarily designed for their type 300 data logger/alarm scanning system but it is also available as a self-contained instrument.

A six-decade digital display on the front of the instrument indicates the time in hours, minutes and seconds and a programme timing switch provides for a contact closure or a voltage-output pulse at 1, 2, 5, 10, 20, 30 or 60-min. intervals. In addition, the same contact closure or voltage pulse can be made available at precise hourly times (up to a total number of selections of 24) by inserting a thin plug into one of the sockets of a 24-hole pin board.

Output signals are available on six of 45 lines, at any one time, to operate a printing device; the output level for the printer being + 12 V, −12 V or −24 V. The programmed output level is −24 V make or break, −12 V make or break, or a + 12 V make or break.

As the basic 50-c/s drive frequency is derived from the normal mains supply, the instrument accuracy depends upon the accuracy of the supply frequency.

Microcell Electronics,
Blackwater,
Camberley, Surrey.

Multi-Speed D.C. Chart Recorder

An improved model of the Rohde and Schwarz Enograph d.c. recorder has recently been announced in the U.K. by Aveley Electric.

The Enograph, a high-speed potentiometric instrument with facilities for linear and logarithmic indication, records all variables which can be represented by a direct voltage between 10 mV approximately and 300 V. Six conventional voltage ranges are provided with input impedances from 5 MΩ for the 1-V range to 50 MΩ for the 300-V range. Also, further ranges are provided, including a 1-V range with the input impedance represented by the insulation resistance of the terminals which can be up to 10^{11} Ω. The input of the instrument is fully floating.

Ten chart speeds for 10 mm/sec to 20 mm/hr are push-button selected; chart speeds are in the ratio of 2 or 3 to 1. Provision is also made for operating the chart from an external drive.

Four types of recording are provided: ink, ball-point pen, electrode and stylus.

Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Solid Tantalum Capacitors

Texas Instruments are now producing a range of solid tantalum electrolytic capacitors in standard-voltage classes of 6, 10, 15, 20, 25 and 35-V d.c. and capacitances from 1 to 330 μF, ± 10% and ± 20% tolerance.

Known as type SCM 'tan-TI-cap' capacitors, they are designed to operate at full voltage rating over the temperature range −55 °C to 85 °C and two-thirds rated voltage up to 120 °C. These capacitors are tested 100% for leakage current stability (noise) and are guaranteed to have a failure rate better than 2×10^{-5} when run at maximum rating.

Other features claimed for these components include sub-miniature industrial standard case sizes, a continuous reverse-capability, and high vibration and shock resistance.

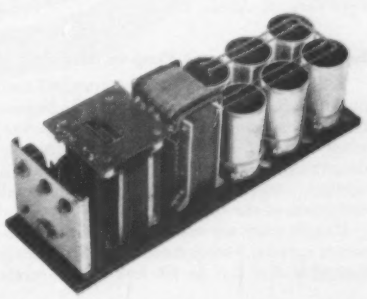
Texas Instruments Ltd.,
Manton Lane, Bedford.

Low-Voltage Stabilized D.C. Supplies

A range of stabilized d.c. power supplies for low voltages has recently been introduced by Claude Lyons.

The series comprises five units: PST-1208 rated at 12 V 8 A, PST-2005 (20 V 5 A), PST-2404 (24 V 4 A), PST-3203 (32 V 3 A) and PST-4802 (48 V 2 A). All models are available for chassis mounting or standard 19-in. rack mounting.

Each supply consists of a conventional d.c. power supply (step-down transformer, bridge rectifier and smoothing), the input to which is fed through a transductor. A transistorized feedback amplifier measures



the output voltage from the supply and continuously adjusts the direct current in the control winding of the transductor to maintain the output voltage constant over a wide range of variation in input (mains) voltage and frequency and in output (load) current.

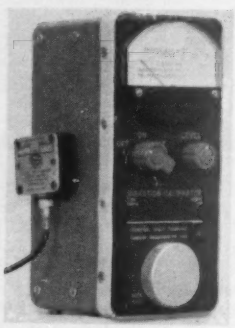
Ripple is of the order of 1% and the output voltage is adjustable over a range of ± 15% about the nominal value.

Claude Lyons Ltd.,
Valley Works, Ware Road,
Hoddesdon, Herts.

Vibration Pickup Calibrator

A compact battery-operated vibration calibrator (type 1557-A), designed for the calibration of accelerometers, vibration pickups, vibration meters, etc., has been announced by the General Radio Company, U.S.A.

Consisting of a transistorized electromechanical oscillator and a battery-operated cylindrical shaker (vibrator), the instrument provides a standard acceleration for 1 *g* (r.m.s.) at 100 c/s. Acceleration output appears at two pillbox-shaped 50-gm discs mounted on an internal cylinder extending through the instrument and projecting at



the sides. The pickup to be calibrated is attached to, or in place of, one of these discs. Acceleration accuracy is ± 10% and frequency accuracy is ± 1%.

With the aid of the built-in panel meter, which indicates the mass of the accelerometer being checked, and the oscillator amplitude control, it is possible to subject pickups of various masses to a standard acceleration of 1 *g* at 100 c/s.

Power for the instrument is supplied by four RM-Y mercury cells whose life is 100 hr.

The total weight of the instrument, including a leather carrying case, is 1·5 kg and the dimensions are 102 × 204 × 102mm

General Radio Company,
West Concord, Mass., U.S.A.

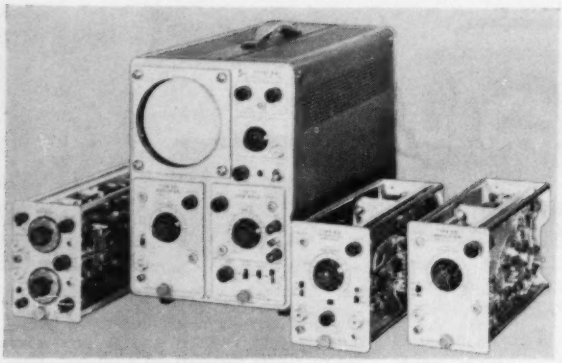
Indicator Units and Plug-In Modules

The Tektronix oscilloscopes types 560 and 561 have been designed as basic indicator units, comprising only the c.r.t. and power supplies, with the overall performance determined by plug-in units; one controlling vertical deflection and one horizontal deflection of the beam.

Plug-in units are available for both instruments varying from a basic amplifier having a passband of d.c. to 400 kc/s with a sensitivity of 1 V/cm to a differential-input amplifier with 1 mV/cm sensitivity. An additional series of plug-in units suitable for the type 561 includes a dual-trace amplifier and a d.c. to 4-Mc/s amplifier.

Distributed in the U.K. by

Livingston Laboratories Ltd.,
*Retcar Street, London, N.*19.



Lissajous Oscilloscope

A compact Lissajous oscilloscope which may be used either as a bench or rack-mounted instrument, and designed to simplify phase-shift measurement and monitoring, has been introduced by North Atlantic Industries, U.S.A.

Two basic models are available and both are being marketed under the trade name 'Liss-a-Scope'.

Model OS-601 is a broad-band instrument designed for operation from 5 c/s to 10 kc/s providing a sensitivity from 5 mV/cm to 300 V/cm in 100 calibrated ranges. Provision is also included for continuous interpolation between ranges without affecting the phase-shift calibration. Matched vertical and horizontal amplifiers permit display magnifications without distortion of at least 10 : 1 over the bandwidth 5 c/s to 50 kc/s and 5 : 1 up to 100 kc/s. Phase-angle resolution is claimed to be better than 0·02°.

A companion model, OS-602, is a modularized ruggedized version of the instrument for single-frequency or narrow-band work.

Electronic Manufacturers Export Corpn,
Plainview, Long Island, N.Y., U.S.A.

Clip-On Volt Ammeter

A new multi-range clip-on volt ammeter, type RS-1, has been announced by Pyramid Instrument Corporation, Lynbrook, N.Y., U.S.A. It is designed for a.c. measurement only and has four current ranges and two voltage ranges covering 0–5 A to 0–100 A and 0–150 V and 0–600 V with an accuracy of 3% f.s.d.

The RS-1 Amprobe relies on magnetic induction for current measurements and can, therefore, be used without cutting conductors; bayonet plug sockets are provided for voltage measurements.

A feature of the meter is the 'Pointer-Lok', which makes it possible to lock the needle in place when taking readings in dark locations.

Distributed in the U.K. by

H. J. Baldwin & Co. Ltd.,
221 *Grand Buildings,*
*Trafalgar Square, London, W.C.*2

Liquid Level Controller

Thomas Industrial Automation have produced a compact conductivity-operated liquid level controller known as the Resistron. It is designed to control the level (in tanks, etc.), of liquids which have a specific resistance of up to 5·0 kΩ/cm³.

The Resistron is available in three versions: type RS500A for high-level control, RS500B for low-level control, and RS500C for continuously-variable sensitivity control and fitted with a high/low level changeover switch.

For control of external circuits, when the liquid level reaches the pre-set point, a relay is provided with one contact normally open and a second contact normally closed, both rated at 5 A 250 V a.c.

The standard version covers mains supplies of 110 V, 240 V and 440 V, 40–60 c/s.

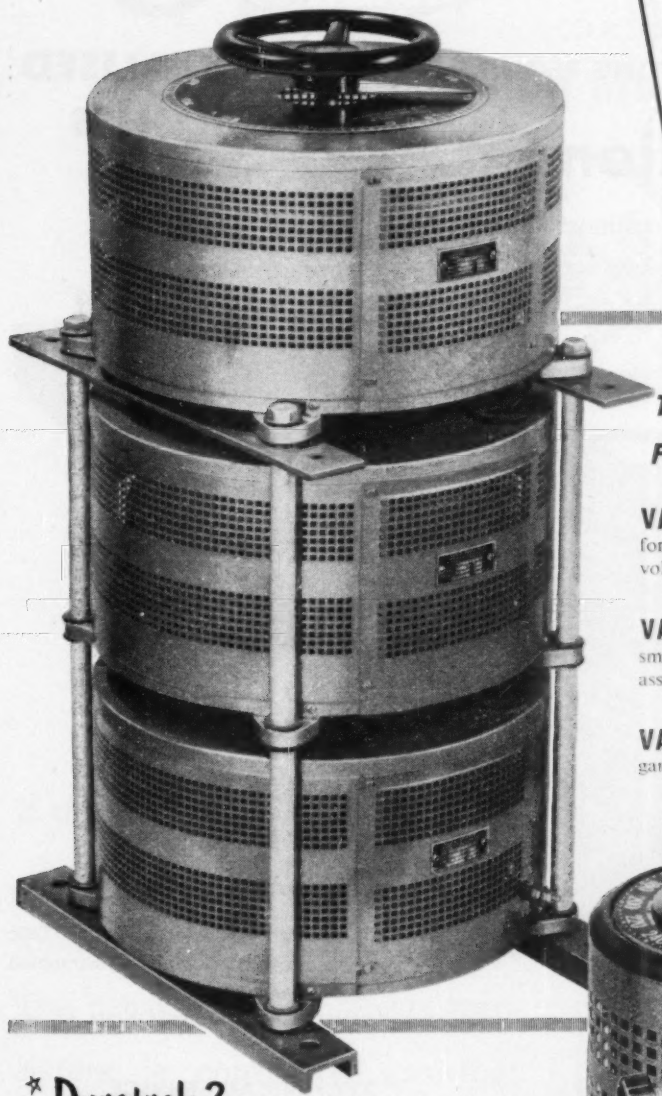
Thomas Industrial Automation Ltd.,
Station Buildings,
Altrincham, Cheshire.

This 3-gang assembly, Type 50-BMG3, will control 22·5 kVA, 3-phase or single-phase according to connection. Larger assemblies can be made.

Variac

Regd. Trademark

WITH **Duratrak** ☆



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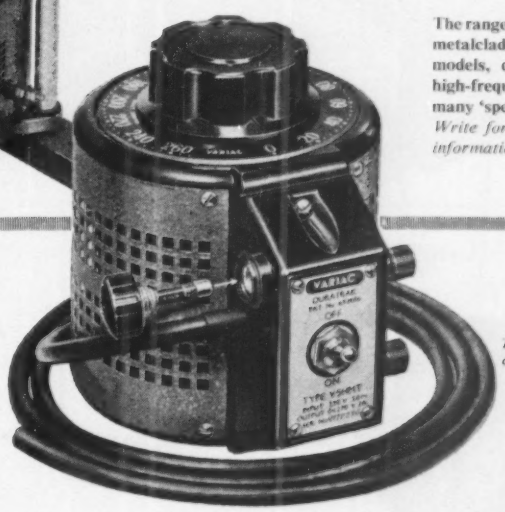
VARIACS are available in a very wide range of models from small units for laboratory and instrument use to large ganged assemblies for three-phase power.

VARIACS are available open or covered, as single units or ganged assemblies, for manual operation or motor-driven.

The range includes portable, metalclad and oil-immersed models, dual-output types, high-frequency types and many 'specials'.
Write for complete information.



This small Variac, Type V-5HMTF, provides an output of 0-270 V, 2 A, from 240 V 50 c/s mains. A still smaller model, Type V-3H is rated at 1 A.

☆ Duratrak ?

Duratrak (Regd. Trademark) — a patented* feature exclusive to Variac — is a special plated contact surface giving longer life, increased overload and surge capacity and maximum economy in maintenance. Duratrak is now standard on all models except Series 50.

*U.K. Pat. No. 693406

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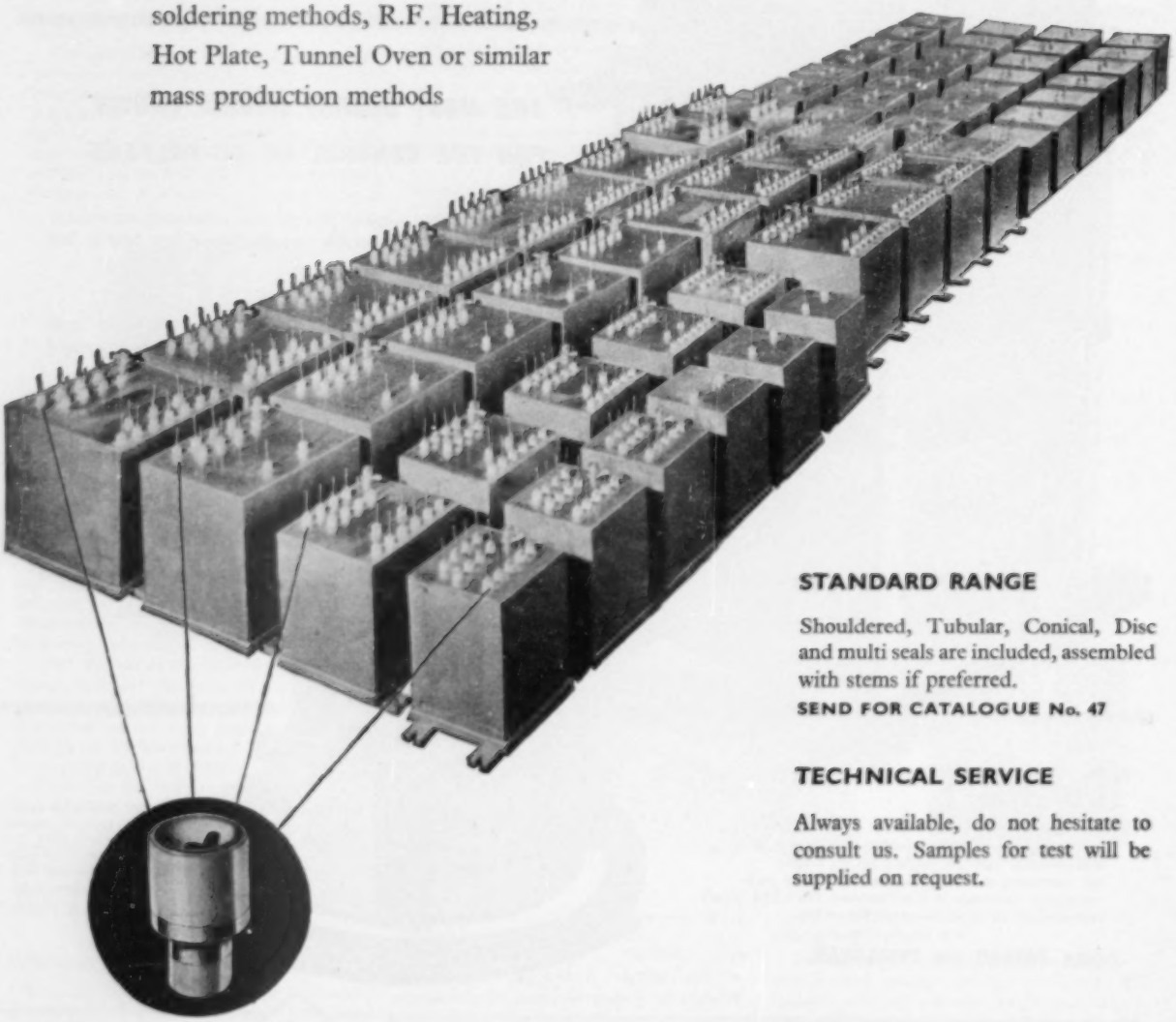

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If you are looking for a VHF pnp drift transistor, the XA131, a recent addition to our range, may be the one you want. It is suitable for use as a **Local Oscillator up to 250 Mc/s or an R.F. Amplifier up to a frequency of 100 Mc/s.**

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Assuming no previous knowledge of the subject, this book deals with the physical processes occurring in transistors, the main emphasis being on the applications of these principles to practical problems of design. The bulk of the book is devoted to the determination of such quantities as input resistance, stage gain, optimum load, power output, values of coupling capacitors and transformer winding inductances. A large number of worked examples are included.

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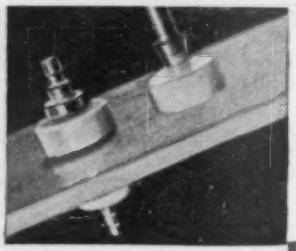

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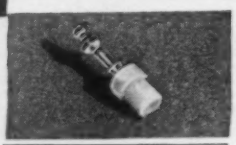

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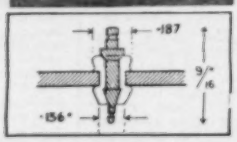

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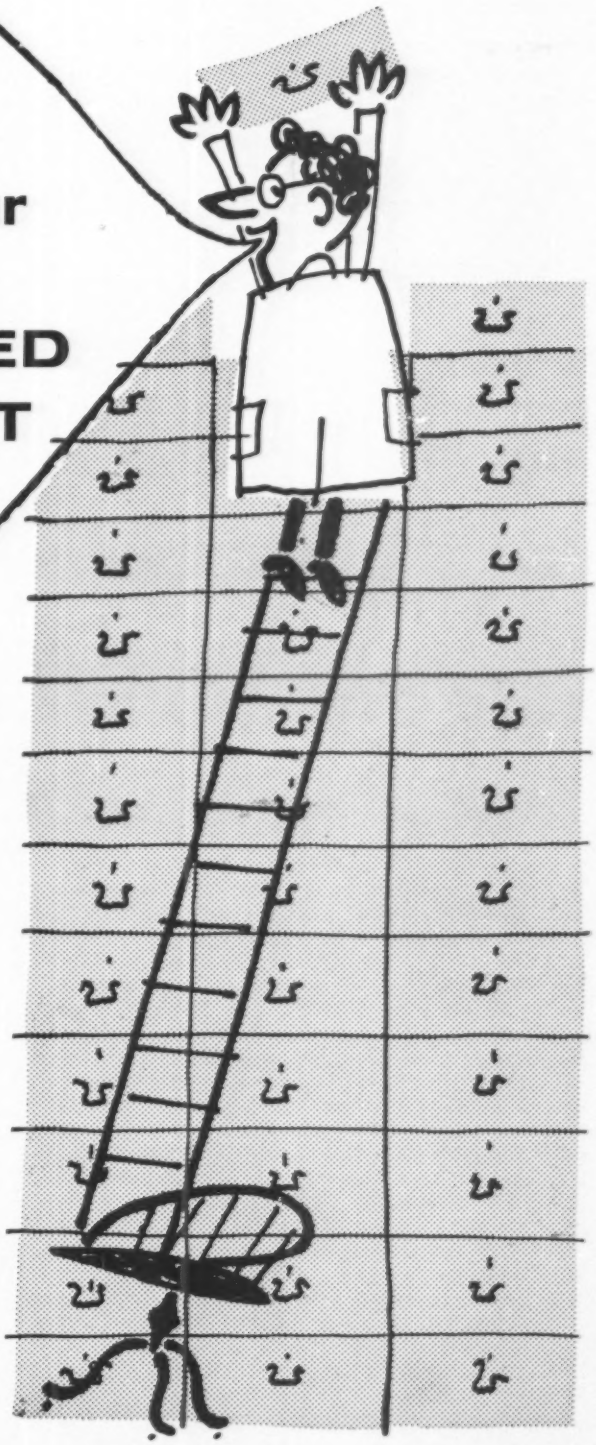

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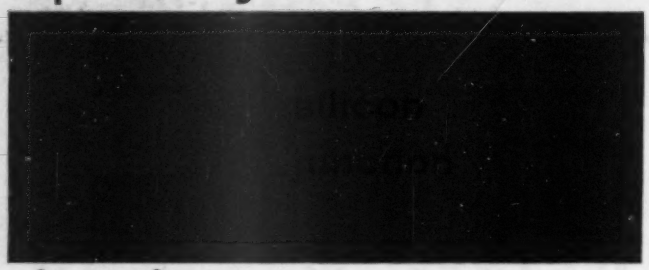

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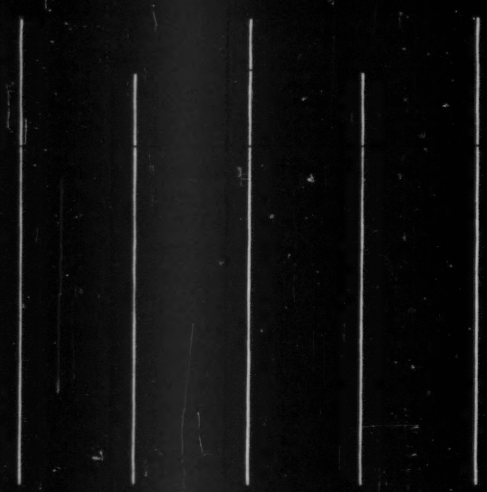

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